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NON-LINEAR DISTORTION PHENOMENA OF MAGNETIC ORIGIN

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Summary. The nature of distortion, and the components of an amplifier in which it can be produced, are discussed. The distortion due to choke coils with iron cores is analysed in detail, and using the results obtained the distortion in amplifier transformers is considered.

Introduction

The subject of amplifiers has been already discussed from various aspects in this Review. In the present article certain characteristics of these units will be analysed in greater detail.

The general purpose of an amplifier is to amplify or magnify an electrical alternating voltage. An amplifier would have an ideal performance if every signal were amplified without any distortion whatsoever, a point which will be considered more closely below. Let us assume that the incoming signal is a certain function of the time, viz., $V_1(t)$; the amplified voltage can generally be represented by another function, $V_2(t)$. Amplification is free from distortion when

$$V_2(t) = c\; V_1(t)$$

for all values of t and any arbitrary form of V; c is here the amplifier gain. In other words, the ratio of the output voltage to the input voltage must remain constant. Deviations from this ideal case can be most easily investigated by assuming that the signal is a pure sinusoidal alternating voltage. The possible deficiencies of an amplifier can be classified under the following heads:

a) The amplifier gain is dependent on the frequency;
b) Phase displacements occur in the amplifier resulting in the input and output signals being out of phase;
c) The amplifier gain varies with the intensity of the incoming signal, with the result that the output voltage is not proportional to the input voltage; this is termed non-linear distortion.

a) This form of distortion is found in every case provided a sufficiently wide frequency range is explored. But it can be always reduced to a satisfactory minimum for any particular frequency range in use.

b) Phase displacement is not an important factor in low-frequency amplifiers, since the ear is unable to distinguish phase displacements in this range. On the other hand, it cannot be neglected in television amplifiers and high-frequency amplifiers in radio receiving sets.

c) Non-linear distortion, which is generally referred to as plain distortion, is obtained in all cases, since amplifiers always contain amplifying valves and transformers which as a rule are made up of non-linear circuits and components. In the present article this particular form of distortion will be discussed.

General Considerations on Distortion

In what respect do distortion phenomena detract from the quality of reproduction? Consider the relationship between the input and output voltages, which are connected by the expression

$$V_2 = f(V_1).$$

If $\quad V_1 = V_0 \sin \omega t$

then $\quad V_2 = f\,(V_0 \sin \omega t).$

While V_2 is thus still periodic, it is in general no longer sinusoidal, although compounded of sine and cosine functions. It may be recalled that Fourier's theorem states that a periodic function can be written as the sum of sine and cosine functions. If, therefore, V_2 is a function of the time, then:

$$V_2(t) = b_0/2 + \sum_{1}^{\infty} (a_n \sin n\omega t + b_n \cos n\omega t) \quad \cdot \quad (1)$$

where a_n and b_n are coefficients given by the equations:

$$a_n = \frac{1}{\pi}\int_{-\pi}^{+\pi} V_2(t)\ \sin\ n\omega t\ \mathrm{d}\ \omega t,$$

$$b_n = \frac{1}{\pi}\int_{-\pi}^{+\pi} V_2(t)\ \cos\ n\omega t\ \mathrm{d}\ \omega t.$$

The output voltage is thus made up of the fundamental frequency and its derived harmonics, so that *non-linear distortion is found to consist in the creation of new frequencies which are absent in the input signal.* The following example will demonstrate this. Assuming that

$$V_2 = AV_1 + BV_1^2 + CV_1^3.$$

then for the case where

$$V_1 = V_0 \cos \omega t:$$

we have:

$$V_2 = AV_0 \cos \omega t + BV_0^2 \cos^2 \omega t + CV_0^3 \cos^3 \omega t,$$
$$V_2 = {}^1/_2\, BV_0^2 + (AV_0 + {}^3/_4\, CV_0^3) \cos \omega t + {}+ {}^1/_2\, BV_0^2 \cos 2\,\omega t + {}^1/_4\, CV_0^3 \cos 3\,\omega t.$$

In most cases the relationship between V_2 and V_1 cannot be expressed by a simple mathematical formula, but can usually be deduced from the form of the curve $V_2(t)$ (cf. *fig. 1*). The following rules are useful in this connection.

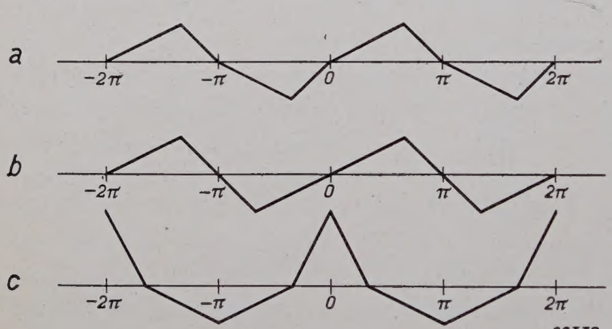

Fig. 1. *a)* Symmetrical function. *b)* Odd function. *c)* Even function.

a) A symmetrical function ($f(t) = -f(t+\pi)$) is composed of the fundamental wave and odd-numbered harmonics.

b) An odd function ($f(t) = -f(-t)$) is composed of sine functions only.

c) An even function ($f(t) = f(-t)$) is composed of cosine functions only.

The first rule in particular has an extremely useful application in amplifier technique for the purpose of reducing the distortion as far as possible. This will be considered in closer detail for a simple amplifying valve by studying the connection between the anode current and the grid voltage (*fig. 2*). The anode current produces a voltage drop at the resistance R which is proportional to the current. Take the point O as the working point. It is seen from the figure how the anode current,

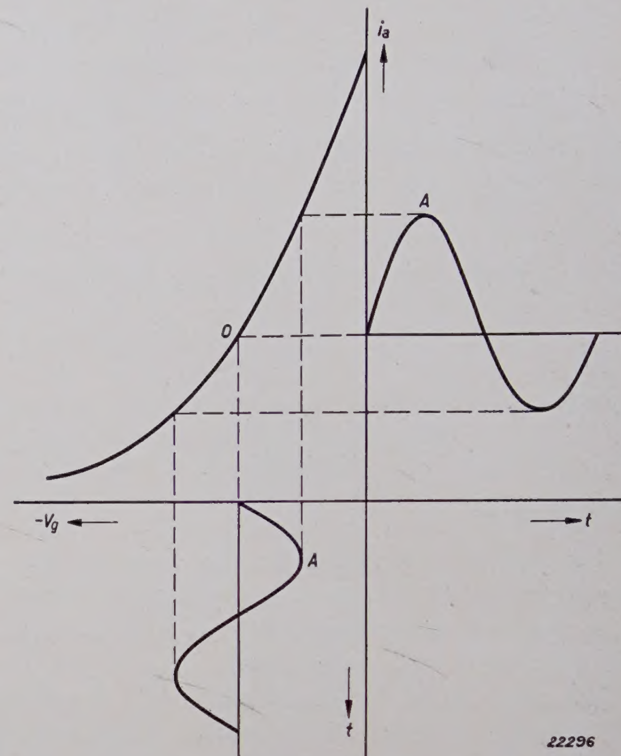

Fig. 2. Anode-current characteristic of an amplifying valve. The anode-current/grid-voltage characteristic is plotted in the top left-hand quadrant. O is the working point. The curve in the bottom left-hand quadrant shows the sinusoidal variation of the grid voltage. In the top right-hand quadrant the anode current is plotted against the time.

and with it the voltage drop at the resistance R, are dependent on the time, when grid voltage has a sinusoidal time function. It is evident that the curve is not symmetrical, for its curvature is much greater at low anode currents than at high currents. If zero time is taken at the point A, the anode current is found to be an even function of the time and on expansion gives a series of cosine functions only.

Now consider two valves connected in a circuit such that the anode current of one valve just reaches its maximum value when the current in the other valve is a minimum. This is the so-called push-pull circuit (see *fig. 3*). We must plot the common characteristic of the two valves. Without entering into details, it may be sufficient to indicate that a symmetrical arrangement is obtained by connecting the valves in push-pull, and in consequence all even-numbered harmonics disappear. In other words: the even-numbered harmonics, which are produced in each of the two valves, are brought into phase opposition so that they compensate each other. Since, in triodes, the second harmonic predominates throughout a wide range over

all other harmonics, the distortion is very considerably reduced in this circuit.

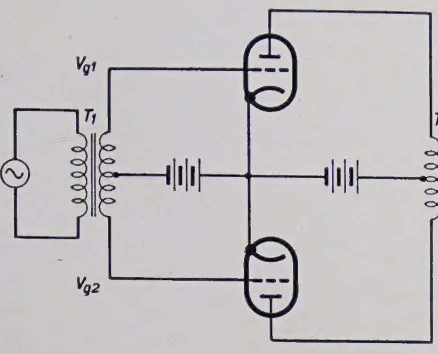

Fig. 3. Push-pull circuit. The transformer T_1 is so wound that V_{g_2} is a minimum when V_{g_1} is a maximum. The anode currents in the two valves are thus in phase opposition; they are combined in the transformer T_2 in such a way that mutual amplification is obtained.

If the input voltage is not sinusoidal, analysis of the conditions becomes more complex. The signal must then be regarded as a sum of sinusoidal oscillations (Fourier series) of different frequencies and amplitudes, and the amplifier has to amplify this sum as a whole. New frequencies are then produced, not only harmonics but also combination tones, i.e. vibrations with a frequency of $m\omega_1 + m\omega_2$, where ω_1 and ω_2 are two input frequencies and m and n are integers. It is just these combination tones which reduce the quality of reproduction, but only in simple cases can their intensity be calculated. One of the principal difficulties is that with two frequencies the distortion is often not the same for both; however, this problem will not be discussed in the present paper.

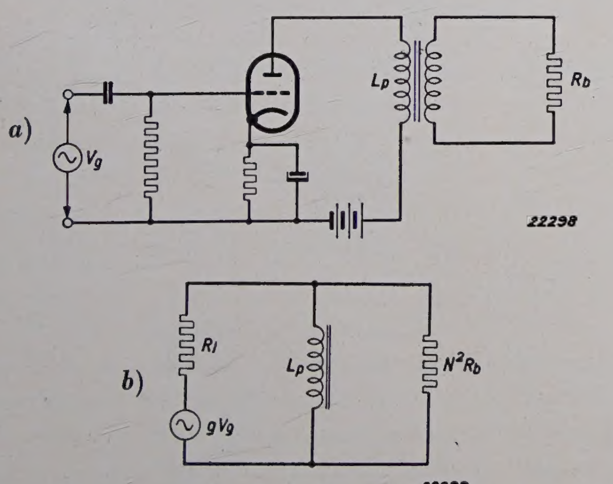

Fig. 4. *a)* Circuit of an output stage. *b)* Equivalent circuit. The amplifying valve has been replaced by the internal resistance R_i in series with the generator gV_g (g = amplification factor of the valve). The transformer with the load resistance has been replaced by the primary self-inductance L_p with N^2 times the value of the load resistance in parallel thereto.

Magnetic Distortions

Those distortion effects will now be discussed in more detail which originate from transformers and choke coils with iron cores present in the circuit. Special reference will be made to choke coils, since transformers and choke coils have the same effect on the distortion phenomena under discussion.

To simplify analysis, we shall consider the distortion produced by an output transformer, and it will be assumed that the load on the amplifier is a resistance and not a loudspeaker as is usually the case (*fig. 4a*). We must first construct the equivalent circuit for this output stage. The amplifying valve can be represented by a current generator in series with the internal resistance of the valve (*fig. 4b*). Since only low frequencies are under consideration (the reasons for this will be discussed later) we can neglect the leakage and capacity characteristics of the transformer. The equivalent circuit of the transformer then consists of the primary self-inductance, which carries a load equal to N^2 times the load resistance (N = transformation ratio of transformer) (see fig. 4). For our purposes this circuit is still comparatively difficult to analyse, so that in the first place we will consider several simpler cases.

Distortion due to non-loaded Iron-core Coils

Undistorted current, i.e. pure sinusoidal current, is passed through the coil; this may be done by connecting a resistance in series with the coil the value of which is very large in comparison with the coil impedance (see *fig. 5*). An e.m.f. appears across

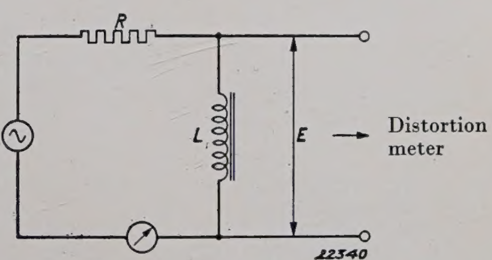

Fig. 5. Distortion measurement in a non-loaded coil. The distortion is eliminated from the current passing through the coil by means of the resistance R ($R \gg \omega L$). The distortion meter places no load on the coil since the input resistance is very high.

the terminals of the coil, and we have to determine its value when the coil has an iron core. The following equations can be employed here:

$$H = \frac{0.4\pi n i}{l} \text{ oersted} \quad \cdots\cdots \quad (2)$$

$$V = n\,O\,\frac{dB}{dt}\,10^{-8}\,\text{volt/cm} \quad \cdots \quad (3)$$

where: H is the magnetic field intensity;
B is the magnetic flux density;
V is the voltage impressed on the coil;
i is the current through the coil;
n is the total number of turns;
l is the mean length of the lines of force, and
O is the cross-section in sq. cm.

Thus already we have an expression connecting the electrical voltage V and the magnetic flux B, and a second expression connecting the electric current i and the magnetic field intensity H. To find the relationship between V and i, we still lack an expression connecting B and H, to deduce which further information must be obtained concerning the magnetic properties of the core iron.

This subject has already been discussed in an article in a previous issue of this Review [1]).

We know that the relationship between B and H:
1) Is not linear,
2) But has an hysteretic character, in other words, a closed curve is traced in the BH plane, when H passes through a cycle (*fig. 6*). The hysteresis loop

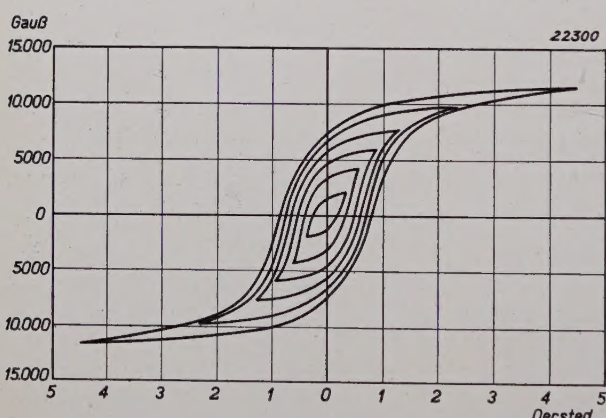

Fig. 6. Examples of hysteresis loops. The various loops relate to different values of B_{max}.

consists of two arms, a rising (lower) branch and a falling (upper) branch; when H increases the ascending curve is traced out and when H diminishes the descending curve is obtained. The form of the hysteresis loop is determined by variations in the magnetic flux density; at very large amplitudes saturation is obtained.

As yet it has not been possible to evolve a mathematical function for the hysteresis loop which is valid for all amplitudes. Only at low flux densities can a relatively simple expression be obtained and to which we shall confine our remarks for the moment. That the curve traced is a loop signifies that energy is lost in the core, these losses being the so-called hysteresis losses; the area enclosed by the loop is a measure of these losses. The form of the loop determines the distortion; if the loop were an ellipse no distortion would occur and merely a phase displacement of B with respect to H. But since the loops have cusps, distortion results. Assume a specific relationship connecting B and H, from which both the hysteresis losses and the distortion can be calculated. It is then found that for many materials the two arms of the loop can be represented by a quadratic equation in terms of the magnetic field intensity, i.e. if we take a new set of bh axes through the lower cusp of the hysteresis loop, the ascending arm is represented by the equation

$$b = \mu_0 h + \nu h^2 \quad \ldots \ldots \quad (4)$$

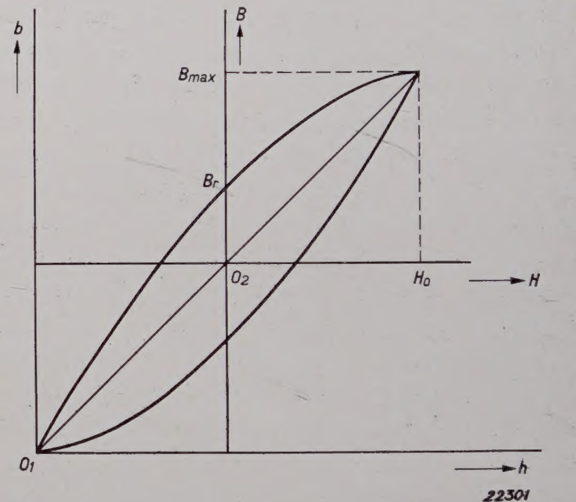

Fig. 7. Hysteresis loop at low amplitudes. The origin of the coordinate axes bO_1h is located at one cusp of the loop; the origin of the axes BO_2H is at the centre of the loop.

where μ and ν are constants of the material (μ_0 is the known initial permeability). This Rayleighs' "law" applies over a wide range about the point $B = H = 0$, i.e. one of the turning points of the loop. We can now deduce directly an expression for the whole of the hysteresis loop by locating the origin at the centre of the loop:

$$B = (\mu_0 + 2\,\nu\, H_0)\, H \pm \nu\, (H_0^2 - H^2), \quad \cdot \quad (5)$$

where $H_0 = H_{max}$, and the plus sign applies to the upper (descending) branch of the loop. From these formulae we get the expressions for the permeability and the remanence, thus:

$$\mu = \frac{B}{H_0} = \mu_0 + 2\,\nu\, H_0 \quad \ldots \quad (6)$$

$$B_r = B_{(H=0)} = \nu\, H_0^2 \quad \ldots \ldots \quad (7)$$

[1]) For a detailed discussion, cf J. L. Snoek, Philips techn. Rev. 2, 77, 1937.

[2]) We here adopt the method of analysis due to H. Jordan, E.N.T, 1, 7, 1924.

From equation (5) connecting the field intensity and the flux density, and equations (2) and (3) connecting the electrical and magnetic magnitudes, we can calculate the voltage appearing across a coil when an undistorted, sinusoidal current: $i = i_0 \cos \omega t$; is passed through the coil. The magnetic field intensity is then, according to equation (2), proportional to i, thus:

$$H = H_0 \cos \omega t,$$

We thus get:

$$B = (\mu_0 + 2\,\nu\, H_0)\, H_0 \cos \omega t \pm H_0^2 \sin^2 \omega t, \quad \cdot \cdot \quad (8)$$

where the minus sign applies for $0 < \omega t < \pi$ and the plus sign for $\pi < \omega t < 2\pi$.

By Fourier analysis we can expand these expressions to a series of sine and cosine functions, thus:

$$B = (\mu_0 + 2\,\nu\, H_0)\, H_0 \cos \omega t + \frac{8}{3\pi}\,\nu\, H_0^2 \sin \omega t -$$

$$-\frac{8}{\pi}\,\nu\, H_0^2 \left\{\frac{\sin 3\omega t}{1{\cdot}3{\cdot}5} + \frac{\sin 5\omega t}{3{\cdot}5{\cdot}7} + \frac{\sin 7\omega t}{5{\cdot}7{\cdot}9} + \ldots\right\} \cdot \quad (9)$$

It is seen that odd-numbered harmonics alone appear in the magnetic induction. This was to be expected since we are dealing here with symmetrical conditions in which even-numbered harmonics disappear. The voltage applied to the coil is found from equation (3) to be:

$$\frac{V}{\omega\, n\, O\, 10^{-8}} =$$

$$= -(\mu_0 + 2\,\nu\, H_0)\, H_0 \sin \omega t + \frac{8}{3\pi}\,\nu\, H_0^2 \cos \omega t -$$

$$-\frac{8}{\pi}\,\nu\, H_0^2 \left\{\frac{\cos 3\omega t}{1{\cdot}5} + \frac{\cos 5\omega t}{3{\cdot}7} + \frac{\cos 7\omega t}{5{\cdot}9} + \ldots\right\} \quad (10)$$

To express the voltage V in terms of the current i_0, we must substitute further in (10):

$$H_0 = \frac{0.4\,\pi\, n\, i_0}{l}.$$

Thus the voltage is made up of the following components:

1) A component with a phase displacement of 90 deg. with reference to the current,
2) A component in phase with the current, and
3) Harmonics.

1) Since $V = L\, di/dt$, where L is the self-inductance, we get for L:

$$L = a\,(\mu_0 + 2\,\nu\, H_0), \quad a = \frac{4\,\pi\, n^2 O}{l}\, 10^{-9} \cdot \quad (11)$$

If the coil has no iron core and no resistance, this component alone is obtained.

2) Since the component is in phase with the current, energy is lost in the coil, viz: hysteresis loss, which may be expressed by saying that the coil has a loss resistance of R_h. This is given by the expression:

$$R_h = a\,\frac{16}{3}\, f\,\nu\, H_0 = \beta\,\frac{16}{3}\, f\,\nu\, i_0, \quad \cdot \cdot \quad (12)$$

$$\beta = \frac{1.6\,\pi^2\, n^3\, O}{l^2}\, 10^{-9} \text{ and } f = \frac{\omega}{2\pi} \cdot$$

where f is the frequency.

3) This component is the distortion required. The amplitudes of the third, fifth and seventh harmonics are given by the expressions:

$$V_3 = \beta\,\frac{16}{5}\, f\,\nu\, i_0^2\,;\; V_5 = \beta\,\frac{16}{21}\, f\,\nu\, i_0^2\,;$$

$$V_7 = \beta\,\frac{16}{45}\, f\,\nu\, i_0^2 \quad \cdot \cdot \cdot \cdot \cdot \cdot \cdot \cdot \cdot \quad (13)$$

It is seen that the amplitude of each harmonic increases with the square of the current. It now remains to compare the distortion voltage with the voltage of the fundamental wave.

If: $\nu\, H_0 \ll \mu_0$, we get:

$$\frac{V_3}{V_1} = \frac{0.2\, n}{l}\,\frac{16}{5}\,\frac{\nu}{\mu_0}\, i_0 \quad \cdot \cdot \cdot \cdot \quad (13a)$$

Hence the percentage distortion is independent of the frequency (the same loop is traced out at all frequencies) and is proportional to the current through the coil. Equation (10) gives for the ratio of the various harmonics:

$$V_3 : V_5 : V_7 = 1/5 : 1/21 : 1/45 \quad \cdot \quad (13b)$$

As an example the results of distortion measurements on a coil composed of nickel-iron sheets are represented in *fig. 8*. The equations satisfactorily

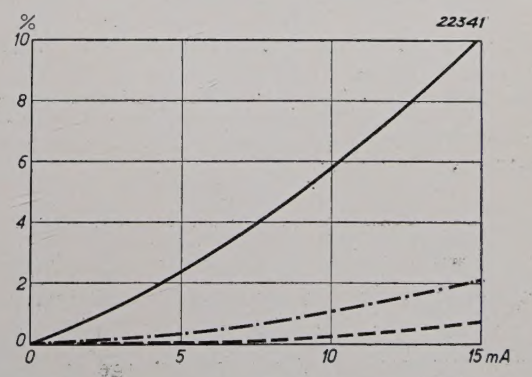

Fig. 8. Distortion in a non-loaded coil with nickel-iron core. ———— third harmonic —·—· fifth harmonic — — — — seventh harmonic.

represent the general shape of the curve, although systematic deviations are found to occur: The distortion is not absolutely proportional to the current, and the ratio of the various harmonics

is not quite constant; the higher harmonics are not as high as found by calculation.

By combining equations (12) and (13) the relationship between the hysteresis resistance and the distortion is obtained, thus:

$$\frac{V_3/i_0^2}{R_h/i_0} = 0.6 \quad \cdot \cdot \cdot \cdot \cdot \cdot \cdot (14)$$

With this expression the distortion of a coil can be calculated when the loss resistance is known. Equation (14), also, is not completely satisfied for this coil; the ratio is not constant, for the distortion does not vary linearly with the current and is smaller than demanded by this equation. On the other hand, it may be concluded from these experiments that in broad outline the theory gives a fairly accurate representation of the distortion at low flux densities.

Distortion in Loaded Coils

In the previous section we have discussed the distortion occurring in non-loaded coils, i.e. in coils which are not shunted by an external impedance. We will now examine the conditions when such an impedance is present, and assume that undistorted current again flows through the coil and impedance, and has such a large amplitude that the current through the coil is the same as in the absence of impedance (see *fig. 9*). Since the

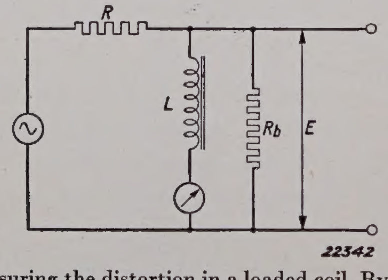

Fig. 9. Measuring the distortion in a loaded coil. By connecting the ammeter in series with the coil, distortion measurements are made for the same values of the fundamental wave as in fig. 4. With high distortion values a selective apparatus must be used, which gives an indication of the amplitude of the fundamental wave alone.

coil is now shunted, the undistorted e.m.f. across the coil will cause a current of the higher harmonic frequencies to flow in the shunt circuit formed by the coil and its external impedance. The flow of this current reduces the distortion voltage in the coil, and the quotient of this voltage drop and the distortion current may be regarded as an internal impedance with respect to the higher harmonics. At a first glance it might be assumed that this impedance has the same expression as applicable to the fundamental wave. But it must be remembered that the coil, which is traversed by both the distortion current and the fundamental wave, is not a linear impedance and in view of this these currents are not independent of each other. The detailed calculations cannot be given here; only the results obtained will be stated.

The impedance for the fundamental wave (i_1) is represented by the expression:

$$Z(i_1) = R_h(i_1) + j\omega L(i_1) \quad \cdot \cdot \cdot (15)$$

where R_h and L are function of i_1; in addition R_h is proportional to the frequency.

For the internal impedance in respect of the third harmonic we then get:

$$Z(i_3) = 26/35\, R_h(i_1) + j \cdot 3\,\omega L(i_1) \quad \cdot (16)$$

It should be pointed out that in this expression the terms R_h and L stand for values related to the amplitude of the fundamental wave. As long as the induction in the iron is low, R_h will remain small (R_h is proportional to the current), so that in the majority of cases we have to take the self-inductance alone into consideration. The same applies for the higher harmonics also. In this way we can calculate the distortion current flowing through the circuit, and hence also the voltage obtained in the internal impedance as well as that applied to the coil. To simplify matters, the external impedance is assumed to be a resistance. The distortion then becomes smaller when the coil carries the resistance as a load. Since the internal impedance increases with the order of the harmonics, the higher harmonics will be reduced to a greater extent than the lower harmonics. Hence, if we make the external resistance equal to ωL for the fundamental wave, the third, fifth, seventh and ninth harmonics will be reduced to:

$$\frac{1}{\sqrt{10}},\ \frac{1}{\sqrt{26}},\ \frac{1}{\sqrt{50}}\ \text{and}\ \frac{1}{\sqrt{82}}$$

of their original values. In practice equation (16) is adequately satisfied. Although it has been deduced for low flux densities (Rayleigh's formula) it is evident that it represents quite satisfactorily the conditions at high flux densities also.

Nevertheless, in certain cases, the distortion may increase when the coil is shunted by an impedance. This may be demonstrated by using as load, a condenser with such a capacity that at the frequency of one of the harmonics it is in resonance with L. The circuit then constitutes a series resonance for this harmonic, and the current flowing through the circuit is limited solely by the loss resistance of the circuit, i.e. by the hysteresis resistance. Hence this current will be very high and in consequence a high voltage will be impressed on the condenser by this harmonic also.

Eddy-Current Coils

In our analysis we have not taken into consideration that eddy-current losses occur frequently in a coil with an iron core. To examine to what extent these currents will modify the results we have obtained, we must again refer to the equivalent circuit of the coil.

We will assume that the frequency is so low that no appreciable displacement of the lines of force as yet occurs as a result of eddy currents, so that the induction is uniformly distributed over the whole cross-section of the iron. In an equivalent circuit, the eddy currents can be represented as flowing in an eddy-current loss resistance. It may be asked whether it is more correct to put this resistance in series or in parallel with the coil. This can be determined by considering the origin of the eddy currents; it cannot be established experimentally, since each circuit can be transformed by calculation into the other. We shall analyse the two forms of circuit arising (see *fig. 10*):

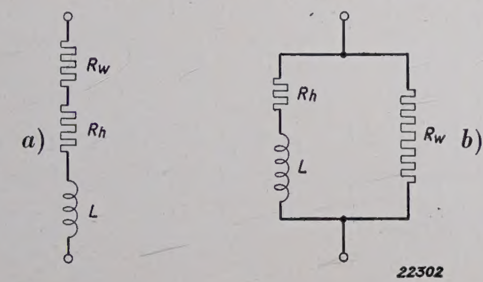

Fig. 10. *a)* Circuit with a coil in series with the eddy-current resistance. L = self-inductance, R_h = hysteresis resistance, R_w = eddy-current resistance.
b) The same circuit with eddy-current resistance in parallel.

1) Eddy-current Resistance in Series.
This circuit does not show that the eddy-current losses are due to currents flowing in the iron, for a circuit of this type generates a magnetic field which is opposed to the initial field. Eddy currents thus reduce the magnetic flux in the core at a given current. This is not indicated in the circuit diagram for the initial current continues to flow through the self-inductance.

2) Eddy-current Resistance in Parallel.
In this arrangement the formation of eddy currents is clearly illustrated. The eddy-current resistance constitutes a load on the coil; if the external current is kept constant the current in the self-inductance must drop. In this circuit, the eddy currents differ in no way from an ordinary load; the circuit is definitely better than the previous one. Furthermore it is an advantage that, contrary to a series resistance whose value increases with the square of the frequency, the parallel resistance is independent of the frequency at low loss values.

The effect of eddy currents on the distortion is now evident. Similar to a resistance due to an external load, these losses reduce the distortion. If the distortion is measured at such a high frequency that the eddy currents can affect the result (their effect increases with the frequency, since the resistance in parallel remains constant), then the value obtained is not the same as the distortion e.m.f. measured without an external load on the coil.

Distortion due to Transformers

After these preliminary remarks, we can return to a discussion of the distortion in an output stage. To make direct use of the above results, we shall make a small alteration in the equivalent circuit (fig. 4b). Previously we substituted for the amplifying valve a voltage source furnishing a voltage $g\ V_g$ and connected it in series with the internal resistance of the valve. The valve can also be replaced by a current source of intensity $S\ V_g$ (S = slope of the amplifying valve) which is connected in parallel with the internal resistance of the valve (*fig. 11*). The circuit is then identical to that shown

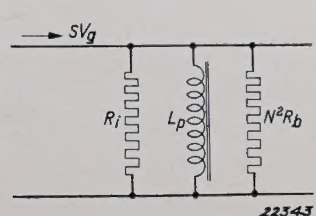

Fig. 11. Equivalent circuit for the output stage. The amplifying valve has been replaced here by a current source SV_g with R_i in parallel.

in fig. 9, when R_i and N^2R_b are combined to form a single resistance. The results arrived at above can then be directly applied; the magnitude of the distortion is determined by the current intensity and by the ratio of the impedance values of the transformer (at different harmonics) to the equivalent resistance of R_i and R_b. It is now seen why a marked distortion has to be expected at low frequencies only. At high frequencies the internal impedance rapidly becomes too great to permit large distortion currents, so that distortion practically disappears; moreover, the current passing through the coil diminishes steadily as the frequency increases.

An amplifier must not only be devoid of distortion but, at a specific low frequency, the amplifier gain must not diminish by more than a certain fraction. Again here the ratio of ωL to $\dfrac{R_iR_b}{R_i + R_b}$ is a determining factor, as may be seen from the following. The gain of the circuit in fig. 11 is given by:

$$V_2/V_1 = \frac{g\,\omega L\,R_b}{\sqrt{R_i^2\,R_b^2 + \omega^2 L^2\,(R_i + R_b)^2}}\,, \quad (17)$$

where V_2 is the voltage applied to R_b and $V_1 = V_g$. The amplification at high frequencies is therefore:

$$V_2/V_1 = \frac{g\,R_b}{R_i + R_b} \quad . \; . \; . \; . \quad (18)$$

If p is the ratio of the gains at low and high frequencies we have:

$$\frac{V_2/V_1(\omega_1)}{V_2/V_1\,(\omega = \infty)} = p\,(\omega_1)\,,$$

so that from (17) and (18) we get:

$$\omega_1 L = \frac{R_i\,R_b}{R_i + R_b}\sqrt{\frac{p^2}{1 - p^2}} \quad . \; . \; . \quad (19)$$

We can now calculate the magnitude of L for the amplifier gain to be reduced to a fraction p at a specific frequency ω_1. It is seen that when

$$\omega_1 L = \frac{R_i\,R_b}{R_i + R_b}$$

the gain drops to 70 per cent of its normal value. Therefore, to determine both the frequency characteristic and the magnitude of the distortion, we must compare the impedance of the transformer with the external load. In some cases the frequency characteristic will be satisfactory, while the distortion will be too high; in other cases the reverse will be found. This will depend entirely on the characteristics of the transformer iron. With $p = 0.7$ the internal impedance with respect to the third harmonic is already three time greater than the external impedance, so that the distortion at this frequency is already considerably reduced. At still lower frequencies, the distortion naturally becomes greater.

In the example dicussed here, a direct current also flows through the coil, which destroys the symmetry of the hysteresis loop, so that even-numbered harmonics also are produced. For the rest the same considerations as given above apply in this case.

Distortion at High Induction Values

The above analysis is only valid for low flux densities. As soon as these latter become large the problem is no longer susceptible to mathematical treatment. Fortunately, the phenomena discussed above then have a more or less similar form, so that it is unnecessary to measure the distortion either for all flux densities or for all practical loads. If it is known how the self-inductance values and the distortion vary with the amplitude in the case of the non-loaded coil, we can estimate the approximate distortion with different loads. However, in this case it is very difficult to take eddy currents into consideration, since in the majority of materials the permeability is dependent on the amplitude. It becomes necessary therefore, to avoid these eddy currents as far as possible, in other words to carry out measurements at low frequencies, especially as the distortion of amplifier coils is only serious at low frequencies.

Finally, *fig. 12* shows as a further example the

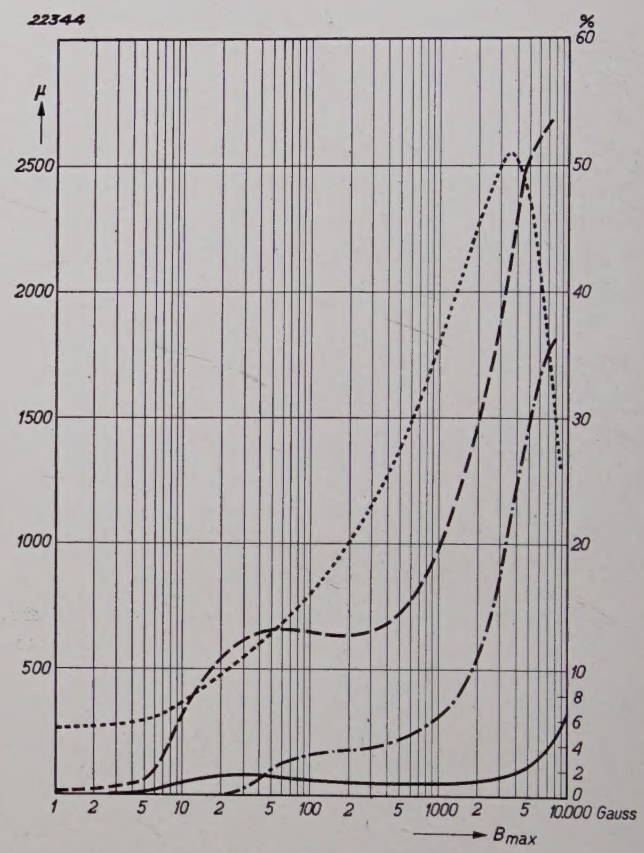

Fig. 12. Distortion of a silicon iron plotted in relation to B_{max}. ...... Permeability, — — — — — Third harmonic on no load. — · — · — · — · — Fifth harmonic on no load. ———— Third harmonic with a load equivalent to a resistance $R_b = \omega L$, where ω is the circular frequency of the fundamental wave.

distortion in silicon iron on no-load, as well as on load, these values being obtained at the fundamental frequency $\omega L = R_b$. It is seen particularly clearly that the internal impedance of the coil satisfactorily fulfils its purpose of reducing the distortion. The continuous curve indicates that these laminations are still effective at $B = 8000$ gauss, with a distortion of less than 5 per cent, and that on the whole the distortion is negligible as compared with that due to the amplifying valve itself.

HIGH-VACUUM GAUGES

by F. M. PENNING.

Summary. The various principles for measuring high vacua are briefly discussed. Some new designs of gauge are described in detail, particularly those which are based on the characteristics of electric gas discharges.

Definition of Gas Pressure; Units

The pressure of a gas is defined as the force applied by the impact of the gaseous molecules against unit surface of the walls of the enclosure containing the gas. The C.G.S. unit in which gas pressures are measured is 1 dyne per sq cm and is termed a microbar [1]). Low pressures are usually expressed in units of millimetres of mercury, a pressure of 1 mm. of Hg being termed the Torricelli or sometimes the Tor. The relationship between these units is:

1 mm Hg = 1 Tor = 1333 microbar = 1333 dynes per sq cm = 1/760 atmosphere.

In this paper, discussion will be confined to gas pressures below one atmosphere, and particularly to those below 1 mm Hg.

Principles of Pressure Measurements

Apart from a direct measurement of the gas pressure, every other property of a gas which varies with the pressure can theoretically be employed as a measure of that pressure. The associated properties which can be used for this purpose are:

1. Gas pressure,
2. Thermal conductivity of the gas,
3. Internal friction or frictional drag,
4. Radiometer effect,
5. Dependent gas discharge, and
6. Independent gas discharge.

Where high-vacuum gauges based on these principles have already been described in the literature (see bibliography), brief mention only will be made to them in the present article; while a more detailed description will be given of any individual new types which have been devised.

Gauges depending on a measurement of the Gas Pressure per se

Manometers or pressure gauges based on this principle obtain a pressure reading by measuring the displacement of the surface of a liquid (liquid gauges) or of an elastic (mechanical gauges). To magnify the sensitivity of measurement, the pressure can be multiplied before measurement, viz, by compression in a known ratio (MacLeod gauge). The surface, which is displaced under the action of the pressure, need not form part of the wall of the actual enclosure, but can be freely suspended. This method is adopted in the

Molecular-Jet gauge (*fig. 1)*, which was designed by Mayer for determining very low vapour pressures [2]). The substance whose vapour pressure is to be determined is contained in the bulb *I*, which is connected to the tubular vessel *II* through a capillary; *II* is highly evacuated through the wide connecting tube *P*. In vessel *II* a small, very light vane with two aluminium discs, 0.01 mm thick and 4 mm in diameter, is suspended; one of the discs is situated directly in front of the opening of *K*. The stream of molecules issuing from *K* displaces the small vane from its position of equilibrium, the magnitude of the angle of deflection being measured by means of a small mirror *S*. In the gauge constructed, this angle of deflection was found to be proportional to the pressure below a limit of 10^{-2} mm. The gauge must be calibrated with a substance of known vapour pressure.

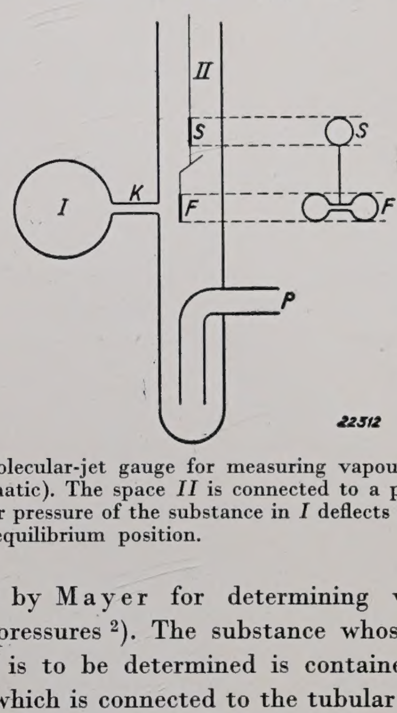

Fig. 1. Molecular-jet gauge for measuring vapour pressures (diagrammatic). The space *II* is connected to a pump at *P*; the vapour pressure of the substance in *I* deflects the vane *F* from its equilibrium position.

1) A considerable divergence of opinion is found in the literature in regard to the units bar and microbar (barye). In Germany and the United States the C.G.S. unit has sometimes been termed the bar. At present the universal definition of these units is: 1 bar = 10^6 dynes per sq cm = 0.99 atmosphere and 1 barye = 1 microbar = 1 dyne per sq cm.

2) H. Mayer, Z. Physik, **67**, 240, 1931.

Mayer obtained a sensitivity of 10^{-6} mm Hg per mm deflection. A serious drawback of this gauge is that it is very susceptible to tremor, and it must, therefore, be carefully set up to protect it against extraneous vibrations. By virtue of its construction this gauge cannot be used for measuring constant gas pressures.

Gauges depending on the Variation of Thermal Conductivity with Pressure

One of the most remarkable achievements of the kinetic theory of gases was to account for the fact that the thermal conductivity of a gas is independent of its pressure. From this it appears impossible to construct a pressure gauge depending for its action on variations in the thermal conductivity. But this conductivity is independent of the pressure only within a narrow range of pressures, for at high pressures the transmission of heat through a gas is a function of the pressure, since in addition to conduction convection of heat also takes place. Theoretically, therefore, a thermal-conductivity gauge is feasible within this range of pressures; a gauge of this type has indeed been described.

At low pressures, the thermal conductivity is independent of the pressure only for as long as the free path λ of the gaseous molecules is small compared with the distance d between the hot and the cold masses ($\lambda \ll d$). It is evident that at zero pressure ($\lambda \gg d$) the thermal conductivity also must be zero and that there must hence be a transition region in which the thermal conductivity varies with the pressure. If $d = 1$ cm, then $\lambda = d$ for a pressure of approximately 0.01 mm Hg and a transition stage becomes feasible between 10^{-1} and 10^{-3} mm Hg.

In this transition region the pressure can be measured as follows: A metal radiator to which a constant quantity of heat is supplied per second is fixed along the centre of a glass tube. The heat loss to the glass wall determines the temperature of the metal, which is hence a measure of the pressure. Gauges of this type require calibration before use, e.g. against a MacLeod gauge. The more common forms of this gauge are:

1) The Pirani Gauge.

In this gauge, a thin wire which is heated electrically and whose resistance is a measure of the temperature is generally used as a metal radiator (*fig. 2a*).

2) The thermo-couple gauge.

In addition to the resistance, other properties of a metal, such as the thermo-electric force, can also be employed for measuring the temperature. In this case a heated wire is fitted with a thermo-couple (*fig. 2b*), and instead of passing an electric

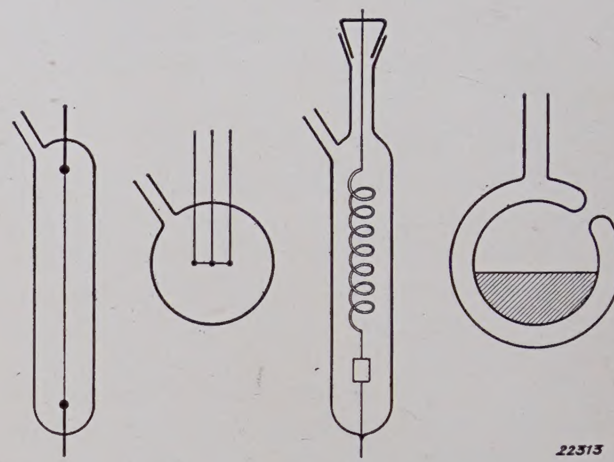

Fig. 2. Principle of the thermal-conductivity gauge.
a) Resistance (Pirani)
b) Thermo-couple (Voege and Rohn).
c) Bimetallic (Klumb and Haase)
d) Dewar flask (Herzog and Scherrer).

current through the wire, heat is supplied to it by irradiation of the thermo-couple with a light-source of constant intensity. In both arrangements the temperature to which the thermo-couple is raised is a measure of the pressure.

3) Bimetallic manometer.

Recently Klumb and Haase [4]) made use of the change in shape of a bimetallic strip (straight strip or helix) to measure the temperature. The application of the principle of a bimetallic helix to the construction of a gauge is shown in *fig. 2c*. The helix is here again heated by the electric current; the rise in temperature causes the small mirror attached to the helix to turn through a specific angle, such rotation being compensated by a torsion head. The number of degrees which the torsion head has to be turned is a measure of the pressure. Results obtained with this gauge for a number of gases are shown in *fig. 3*. The time taken to obtain a reading with the gauge is comparatively long, being about 1 minute.

4) Thermal-conductivity gauge depending on the Rate of Volatilisation of Solid or Liquid Substances.

A gauge constructed on this principle has been described by Herzog and Scherrer [5]). The vessel in which the pressure has to be measured communicates with the space between the walls of a

[3]) Th. Haase, G. Klages and H. Klumb, Phys. Z., **37**, 441, 1936.

[4]) H. Klumb and Th. Haase, Phys. Z., **37**, 27, 1936.

[5]) G. Herzog and P. Scherrer, Helv. Phys. Acta, **6**, 277, 1933.

double-walled sphere which is filled with solid carbon dioxide (*fig. 2d*). The sphere is thus a type

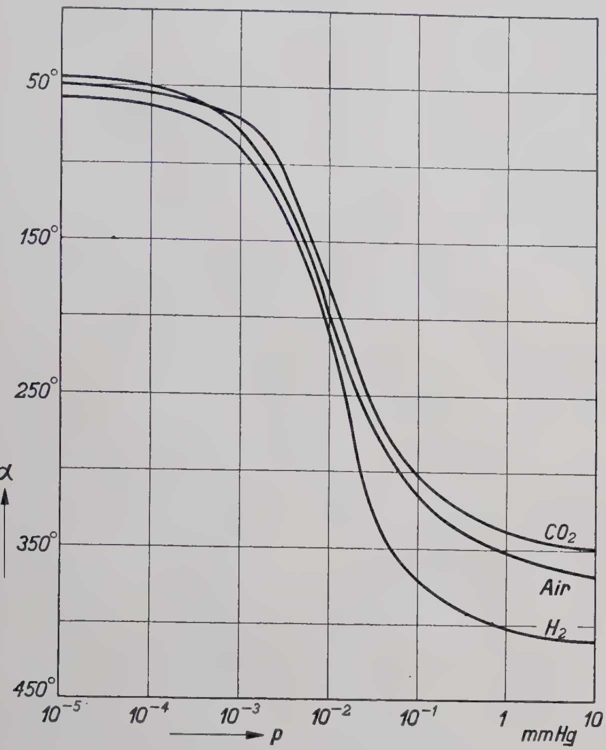

Fig. 3. Angle α through which the torsion head of the bimetallic manometer must be turned to compensate the rotation of the mirror.

of Dewar flask in which the carbon dioxide volatilises the slower the lower the pressure of the gas between the walls of the vessel, so that the rate of volatilisation of the carbon dioxide is a direct measure of the pressure. This gauge is not intended for accurate pressure measurements.

Gauges depending on the variation of Internal Friction with Pressure

Similarly to the thermal conductivity, the internal friction of a gas is also independent of the pressure as long as $\lambda \ll d$, but over a specific range of pressures this property also can be used for measuring high

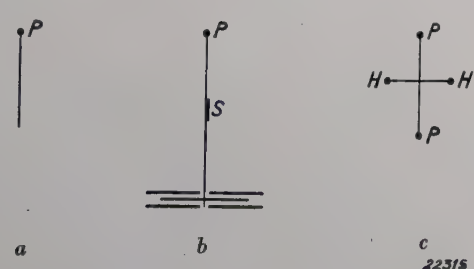

Fig. 4. Principles of the damping-type gauges.
a) Oscillating quartz fibre.
b) Rotatable disc with small mirror *S*.
c) Rotatable quartz cross *HH*; the surface of *H* is used as a convex mirror. The point of suspension is marked *P* in each case.

vacua. Gauges operating on this principle again require special calibration, while the quantitative readings obtained depend on the type of gas under measurement. There are two types of this gauge, viz:
1) Damping gauges, in which the damping of linear or circular oscillations is measured. Of the three forms which have been devised, the first has a thin fibre of glass or other material which is set in vibration by an impulse (*fig. 4a*), while in the second and third forms circular oscillations are imparted by small magnets to an oscillator, either a disc (*fig. 4b*) or a small quartz cross (*fig. 4c*), suspended by a fibre; in the latter case one of the small spheres on the cross acts as a mirror. In all three forms the damping of the motion is a measure of the pressure.
2) Molecular gauge of Langmuir and Dushman in which a rapidly-rotating aluminium disc *A* applies a torque to a thin mica disc *M* (*fig. 5*), this torque being directly proportional to the pressure of gas in the gauge.

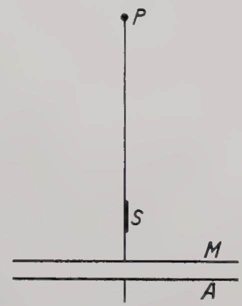

Fig. 5. Molecular gauge of Langmuir and Dushman. *A* - Rapidly-rotating aluminium disc; *M* - mica disc, whose deflection from the equilibrium position is read off by means of the small mirror *S*.

Gauges depending on the Radiometer Effect

These gauges are based on the principle that in very highly-rarefied gases (free path » dimensions of the containing vessel) two surfaces at different temperatures exert a mutual mechanical force. The rotation of the vanes in the Crookes' radiometer also is due to this force. Consider a flat plate *B* (*fig. 6*) at the absolute temperature *T*

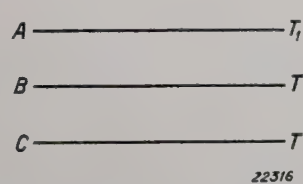

Fig. 6. Mechanical force applied to a plate *B* at the temperature *T* situated between a second plate at the same temperature and a third plate at a higher temperature T_1. Owing to the higher velocity of the molecules arriving from above, the pressure applied to the upper side of *B* is greater than that applied to the under side.

which is located between two other plates A and C at temperatures of T_1 and T respectively. At equilibrium the number of molecules striking against each sq cm of the three plates will be the same for each plate, viz, n. At the low gas pressures in question here nearly every molecule which strikes against B, will have registered its last impact at the surface of the opposite plate and will therefore be at the temperature of this plate. Molecules at a temperature of T_1 thus strike against the upper surface of B and molecules at a temperature of T against the lower surface; the molecules arriving from B are also at a temperature of T. The force applied to B per sq cm is, according to the kinetic theory of gases, $nm\ (v_a + v_b)$, where m is the molecular weight of the gas in question, v_a and v_b the average velocity components perpendicular to the surface with which the gaseous molecules either arrive at or leave the surface respectively, and n is the number of collisions per sq cm per sec. Since v is proportional to $\sqrt{T}$, i.e. $v = c\sqrt{T}$, the difference in pressure p between the top and bottom surface of B is:

$$\Delta p = m\, n\, c\ (\sqrt{T_1} - \sqrt{T}).$$

If the external wall of the enclosure is at the temperature T the pressure in the rest of the enclosure will be

$$p = 2\, m\, n\, c\ \sqrt{T}$$

and hence

$$\Delta p = \frac{p}{2}\left(\frac{\sqrt{T_1}}{\sqrt{T}} - 1\right).$$

The resultant force is, therefore, independent of the type of gas in the gauge.

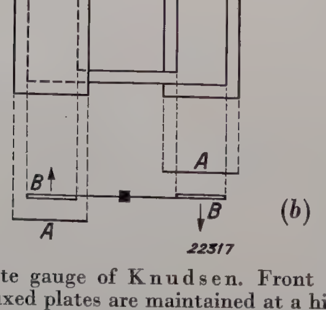

Fig. 7. Absolute gauge of Knudsen. Front view (*a*) and plan (*b*). The fixed plates are maintained at a higher temperature than the rest of the enclosure and thus apply a force on the vanes B in the direction of the arrow. The rotation of B is read off by means of the small mirror S.

1) Absolute gauge of Knudsen.

The absolute gauge of Knudsen consists of the type in fig. 6, but with a glass wall in place of the plate C. The two strips B (*fig. 7*) have been combined to form a small frame which is suspended by a fibre F. The fixed plates A are electrically heated to a temperature T_1, which is higher than the rest of the enclosure, and thus apply a pressure on B in the direction of arrow; the rotation of B which is read off by means of a small mirror attached to F, is a measure of the pressure. The absolute value of the pressure is obtained from the dimensions and the period of oscillation, etc., of the system using the equation given above. The Knudsen gauge is the only gauge which always give a direct reading of the absolute pressure of a gas; this is only possible with the Mac Leod gauge when the gases and vapours present obey Boyle's law.

A short time ago Du Mond and Pickels [6]) described a new form of this gauge, which is suitable for general purposes. In this design a scale is provided on the gauge, which permits a direct reading of the pressure between 10^{-4} and 10^{-6} mm Hg. This particular gauge also requires calibration, and still shares the same disadvantages as the original Knudsen manometer in that it needs careful handling, is difficult to make and can only be used when suitably protected against vibration.

2) Aluminium-Foil gauge of Knudsen.

This gauge is based still more closely on the principle illustrated in fig. 6. A vertically-suspended aluminium foil of the type used in electroscopes here serves as the plate B, its deflection from the equilibrium position being a measure of the pressure.

3) The Molecular vacuum-meter of Gaede [7])

A new gauge of this name and based on the principle under discussion here was recently placed on the market. Two strips HH mounted along the glass wall (*fig. 8*) which are raised to a suitable temperature by electrical heating elements here act as the hot plates. A small frame with two thin strips BB is suspended from a fibre. The equilibrium position which BB assumes under the action of the gas pressure lies in the plane EE and is made to coincide with the elastic equilibrium position. To displace the frame BB, e.g. to the position shown in fig. 8, a specific force which is directly proportional to the pressure must be applied to it. This deflection is obtained by passing a current i through

[6]) J. W. M. Du Mond and W. M. Pickels, Rev. Sci. Instr., **6**, 362, 1935.

[7]) W. Gaede, Z. techn. Phys., **15**, 664, 1934.

the coils S, thus causing a rotation of the small magnet M suspended from the fibre. The magnitude of i for a particular position of BB is, therefore, a measure of the pressure, which latter is read

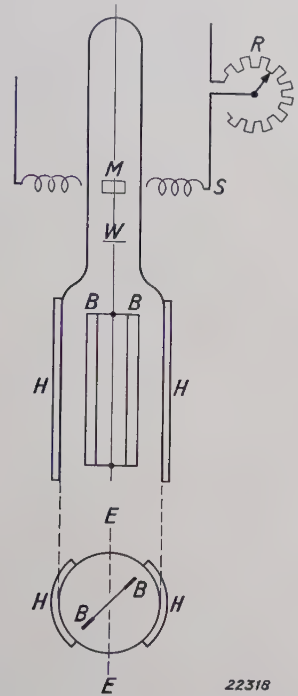

Fig. 8. Molecular vacuum-meter of Gaede. The strips H on the glass wall are raised to a temperature higher than the rest of the enclosure by means of heating elements placed in contact with them. A thermal directive force is applied to the small frame BB provided it is not kept in its equilibrium position EE by an elastic opposing force. By means of the small magnet M and the current i in the coils S, the frame is brought back to its initial position. i is proportional to the pressure in the tube. The magnitude of i which is read off from the adjustment of the variable resistance is a measure of the pressure.

off directly on the control knob of the variable R. The position of BB is adjusted by means of the pointer W. This gauge is calibrated empirically, although the deflection is independent of the type of gas used. Since the "thermal" directive force determines the period of oscillation of the system as well, the latter can also be taken as a measure of the pressure; the relative accuracy of this method is stated to be one per cent. The gauge can be used as a damping manometer also (see above), in which case readings are governed by the internal friction, i.e. to a first approximation by the molecular weight of the gas present. By combining the two methods, the molecular weight can be determined in addition to the pressure.

Gauges based on a Dependent Gas Discharge

In the case of a dependent gas discharge the primary carriers of electricity are not produced by the discharge itself, but derived from an external source, e.g. from a hot wire or by an actinic action. If the electrons, ions or radiation quanta of sufficient energy generated in this way pass through a gas the latter will be ionised; if the pressure of the gas is sufficiently low the number of ions and electrons produced will be proportional to the pressure and the path traversed by the ionised particles. *Fig. 9* shows the number of ionisations N

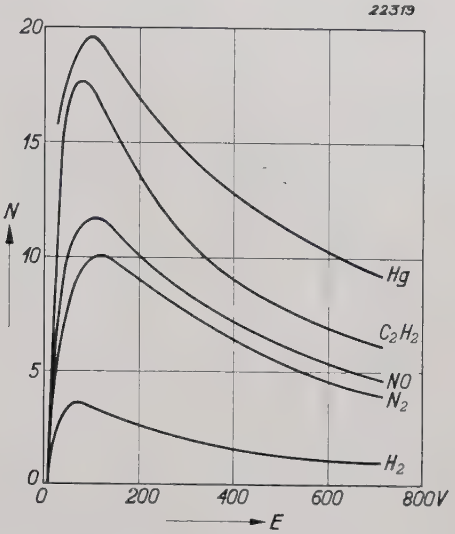

Fig. 9. Ionisation by electrons plotted against the accelerating potential E in volts. N = number of ionisation reactions per cm of path and per mm pressure.

per cm of path and per mm of pressure of the electrons in a variety of gases as a function of the electron energy. It is seen that for most gases N is a maximum at an energy of the order of 100 electron-volts, and that therefore for measuring low pressures a mean electron energy of this magnitude should be used. Since a much smaller number of ionisation reactions is produced by positive ions and radiation quanta of moderate energy, the importance of using electrons for ionisation is evident.

Ionisation Gauge.

In the gauge of this name, ionisation is produced by electrons emitted from a hot tungsten filament F (*fig. 10*). The apparatus is generally in the form of a triode, in which the grid G and the anode A are concentric with the heated filament F. One of the electrodes A or G is maintained at a negative potential with respect to the filament, so that ions alone flow towards it (current i_+) while the primary and secondary electrons can move only towards the third and positive electrode (current i_-). As long as the free path of the electrons is large compared to the path traversed, the pressure is then proportional to i_+/i_-. If i_- is made large, e.g. 10 milliamps, and a sensitive galvanometer is used for measuring i_+, pressures down to 10^{-6} mm can be

measured. If desirable, the ion current also can be amplified to facilitate measurements.

The maximum sensitivity with the method is obtained when G is made positive, for part of the

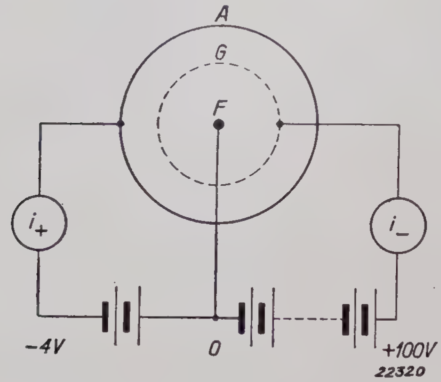

Fig. 10. Ionisation gauge. The ion current i_+ flows to A, and the electron current i_- to G. The pressure is proportional to i_+/i_-.

electrons will then shoot through the mesh of the grid G and describe longer paths before coming back to G, thus having more chance to cause ionisation. By means of a magnetic field the path of the electrons can be still further lengthened [7]).

It follows from fig. 9 that the readings of the ionisation gauge depend on the type of gas used. One of the first observations of Found and Dushmann was that the ionisation was proportional to the molecular weight of the gas used, but it is seen from fig. 9 that this is not the case, and more accurate measurements carried out later by Reynolds [8]) with ionisation gauges showed that this proportionality is not generally valid.

Instead of by direct measurement of the ion current, the pressure of a gas can be determined in still another way with the ionisation gauge. If A is taken as the anode and G as the negative electrode, the insertion of a resistance in the grid circuit will not affect the anode current in any way if the electrodes are located in a vacuum, since no grid current will flow. When gas is admitted to the gauge a positive ion current will flow towards the grid, and the insertion of a resistance in the path of this current will alter the grid voltage and in consequence the anode current also. The alteration in current may be used as a measure of the pressure [9]). Another feasible method is to use the ion current for charging a condenser which is connected either to the grid of another vacuum triode or to a relay-amplifying valve so that the anode current of the second valve is dependent on the condenser voltage [10]). The total rate of change of anode current of this second valve is then a measure of the pressure.

Finally, instead of deriving an electron saturation current from the cathode, the anode voltage can be made so small that the current in vacuo is limited by the electron space charge. If positive ions are then generated at low gas pressures, these are able to neutralise the space charge due to a larger number of electrons and thus allow the cathode current to increase considerably. In this method of measurement proposed by Spiwak and Ignatow [11]) the anode current itself is a measure of the pressure. Theoretically this method was already employed some years previously by G. Hertz [12]) for measuring the ionisation potential of gases.

Gauges based on an Independent Gas Discharge.

In an independent gas discharge, the particles responsible for carrying the charge are generated in the gas or at the electrodes by the discharge itself and without any external agency other than the applied voltage. Although in this case a number of factors, such as the current, current density, and the dimensions of the various components of the discharge, are dependent on the pressure, these have hitherto been used rarely for quantitative pressure measurements, since in general they are difficult to reproduce and also vary within wide limits with the characteristics of the gas present [13]). In order to obtain a rough idea of the vacuum in a glass vessel, a common method used in high-vacuum technique is to impose a high-frequency electrical field and observe the luminous effect resulting therefrom. Down to about 10^{-3} mm Hg the gas itself then appears luminous, while at lower pressures the glass walls alone exhibit a green fluorescence.

This form of gas discharge ceases at a gas pressure of approximately 10^{-3} mm, when using a direct voltage or low-frequency alternating voltage in a valve of moderate dimensions and at pressures of the order of 1000 volts, and for this reason will be quite useless for making measurements at low pressures. But by using a magnetic field and suitably disposing the electrodes, measurements

[8]) N. B. Reynolds, Physics, **1**, 182, 1931.

[9]) H. Teichmann, Z. techn. Phys., **9**, 22, 1928.

[10]) R. Sewig, Z. techn. Phys., **12**, 218, 1931.
A. Butschinsky, Techn. Phys. U.S.S.R., **3**, 223, 1936.

[11]) A. Spiwak and A. S. Ignatow, Sow. Phys., **6**, 53, 1934.

[12]) G. Hertz, Z. Phys., **18**, 307, 1923.

[13]) By using high alternating voltages (7000 volts eff.) Wellauer obtained good results in a range of pressures from $5 \cdot 10^{-3}$ to 10^{-1} mm Hg (Arch. f. Elektrot., **24**, 4, 1930).

can be made at much lower pressures and on this principle a simple gauge can be constructed for pressures down to 10^{-5} mm.

Philips' Vacuum-meter

The Philips' vacuum-meter is based on the principle just discussed and will be described in greater detail by reference to *fig. 11*. If two plates

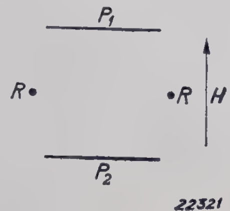

Fig. 11. Principle of the Philips' Vacuum-meter. The electrons at the plates P_1 and P_2 (connected together as a common cathode) do not move along straight lines to the ring RR (anode), but oscillate in spiral paths between P_1 and P_2 under the influence of the magnetic field.

P_1 and P_2 situated opposite to each other form the cathode for the discharge and a ring R acts as a anode, then in the absence of an applied magnetic field the electrons emitted from P_1 and P_2 will travel along curved paths to the ring R. But if a sufficiently powerful magnetic field is applied in the direction perpendicular to P_1, the electrons emitted from P_1 will travel in narrow spirals round the magnetic lines of force towards P_2, will be repelled towards P_1 by the retarding electric field and will thus oscillate many times between P_1 and P_2 before reaching the anode. In consequence, even at very low gas pressures the electrons will be able to collide with a sufficient number of gas molecules producing an adequate ionisation to permit an independent discharge through the gas.

The gauge constructed on this principle is shown in *fig. 12*. The plates P and the ring R are located

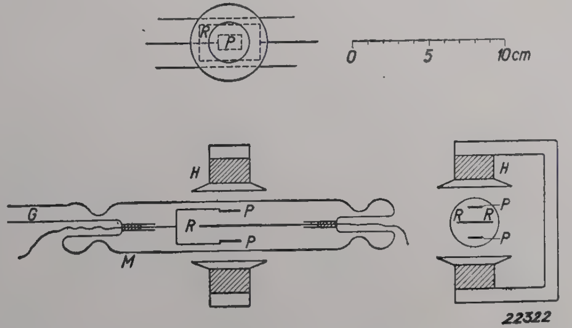

Fig. 12. Philips' Vacuum-meter with two plate-type cathodes P_1 and P_2 and a ring anode R in the field of a permanent magnet H.

in the field of a permanent magnet H, which midway under the pole pieces has a field intensity of approximately 370 oersted. The electrodes are connected up according to the circuit shown in *fig. 13*, viz, through a resistance of 1 megohm to a low current source of 2000 volts. The current i through the gauge

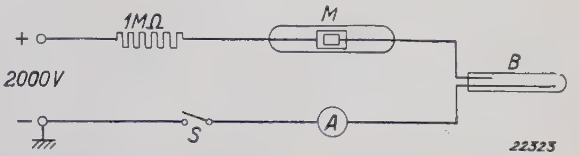

Fig. 13. Circuit of the Philips' Vacuum-meter with the tuning lamp B connected as a current measurer, and the micro-ammeter A.

valve M is hence a measure of the gas pressure, and can be read off on a micro-ammeter with resistances connected in parallel. In qualitative measurements, the instrument can be replaced by a small glow-discharge lamp, such as the Philips 4662 tuning lamp in which the length l of the glow discharge is a measure of the current flowing and hence also of the pressure in the valve M.

The relationship between i or l and the pressure p is shown in *fig. 14* for a pressure range from $2 \cdot 10^{-3}$ to 10^{-5} mm Hg. The curves are based on mean

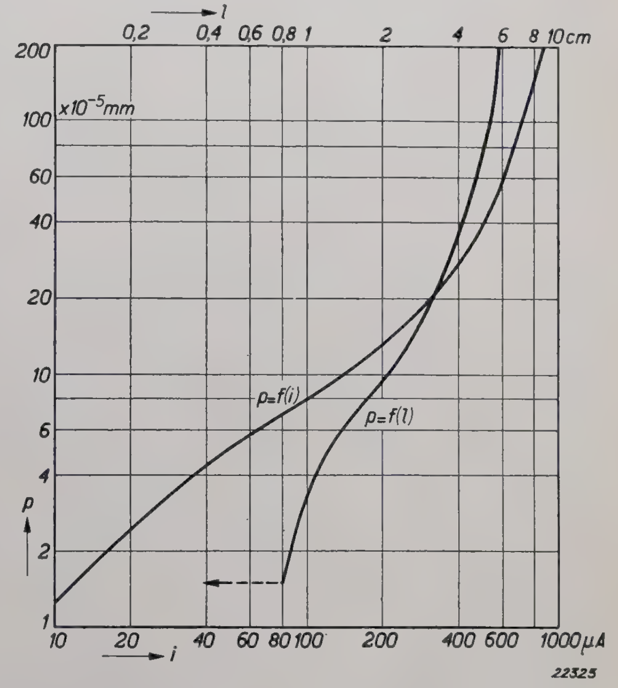

Fig. 14. Gas pressure p in the Philips' Vacuum-meter plotted as a function of the discharge current i or the length l of the glow-discharge in the tuning lamp (mean values for H_2, CO, A and air).

values for air, hydrogen, carbon oxide and argon, while the values of p are in agreement within a factor of 2. This gauge [14]) which has been described in detail in another paper [15]) has the advantage of being extremely simple in construction and does

[14]) Marketed by E. Leybold's Nachf. A.G., Cologne-Bayental.
[15]) F. M. Penning, Physica, 4, 71, 1937.

not require protection against tremor, while one reading made without preliminary adjustment is sufficient to measure a pressure value. It is therefore extremely suitable for industrial use and permits simultaneous and continuous supervision of a multiplicity of pumps. Moreover, it can be sealed to the apparatus of vessels being evacuated or in close proximity to them. Compared with the Mac Leod gauge, it has the advantage of giving continuous and instantaneous readings which are easily visible at a distance; it can be used also for measuring the pressure of condensible vapours and does not itself introduce a supplementary vapour pressure. The latter feature is important when the gauge is used for measurements in conjunction with oil high-vacuum pumps. In conclusion, it should be stated that in the first models of this gauge made the cathode plates P were of iron, but it was shown by experience that the sensitivity of the gauge was increased when the cathode was made of thorium or zirconium; moreover with these metals cathode sputtering was less than with iron.

Tube connecting Gauge and Gas Chamber

In gauges in which gas is continuously consumed, as for instance in those instruments designed on the electrical-conductivity principle, a pressure gradient is produced in the tube connecting the gauge and the gas chamber; this tube must therefore not be made too narrow.

Other potential errors in pressure measurement may be caused by inserting a liquid-air trap between the gauge and the gas chamber, when, in particular, the pressure of condensible gases and vapours, which are given off in the gas chamber, escapes measurement entirely or is only partially measured. A second source of error may be caused by the bore of the trap *not being the same where it enters or leaves the liquid air* [16]). In *fig. 15* the section between A and B represents a trap at a temperature T', with one end of diameter d and the other of diameter D; to the left of A and the right of B the temperature is T. The pressure to the left of A is p, and that to the right of B is P. Over the pressure range in which the conditions of flow conform with the Poiseuille formula (free path $\lambda \ll$ diameters of tubes d and D) we have:

$$\lambda \ll d,\ p = p' = P.$$

[16]) M. Rusch and O. Bunge, Z. techn. Phys., **13**, 77, 1932.

In the range covered by the laws of Knudsen ($\lambda \gg D$) we have:

$$\lambda \gg D,\ \frac{p}{p'} = \sqrt{\frac{T}{T'}} \text{ and } \frac{P}{p'} = \sqrt{\frac{T}{T'}}, \text{ also } p = P.$$

Although d and D are different and no errors are likely to accrue in the two boundary pressure ranges, this is no longer true for the transition range in which λ is approximately equal

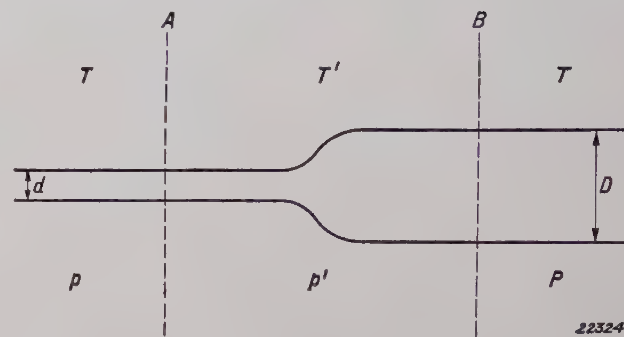

Fig. 15. Gas pressure in a tube of variable diameter at different temperatures.

to d. In this range p/p' and P/p' are determined by d and D respectively, and for $d \neq D$ we have also $p/P \neq 1$; errors of 10 per cent can readily accrue here. If a trap is necessary, then if accurate measurements are required one with a uniform cross-section should be used.

Synopsis

A synopsis of the more common high-vacuum gauges in use is given in Table I, together with the principal ranges of pressures for which they are suitable.

BIBLIOGRAPHY.

S. Dushman. Production and measurement of high vacua. Schenectady, 1922.

A. Goetz: Physik und Technik des Hochvakuums. Brunswick, 1926.

F. H. Newman: The production and measurement of low pressures. London, 1925.

G. W. C. Kaye: High vacua. London, 1927.

S. Dushman: Recent advances in the production and measurement of high vacua. Journ. Franklin Inst. **211**, 689, 1931.

Table I.

Type of Gauge	Range (mm Hg)											Pressure reading dependent on type of gas	Principle
	10^{2}	10^{1}	10^{0}	10^{-1}	10^{-2}	10^{-3}	10^{-4}	10^{-5}	10^{-6}	10^{-7}			
Hg gauge	—	—	—	—								No.	Pressure
Macleod gauge			—	—	—	—	—	—				Partially*)	Press. after preliminary compr.
Pirani gauge (fig. 2a)					—	—	—					Yes	Thermal conductivity.
Knudsen gauge (fig. 7)						–	—	—	—			No	Radiometer effect.
Molecular vacuum-meter (fig.8)			—	—	—	—	—	—	—	—		No	Radiometer effect
Ionisation gauge (fig. 10)						–	—	—	—			Yes	Ionisation by electrons
Philips' vacuum-meter (fig. 11)							—	—				Yes	Gas discharge in magnetic field

*) In the presence of condensable vapours.

THE USE OF AMPLIFIERS (REPEATERS) IN TELEPHONY

by W. SIX and H. MULDERS.

Summary. Following a discussion of the principle of two-wire and four-wire repeaters for amplifying speech currents in telephony, three different types of repeater are described, viz., a four-wire repeater equipped with triodes, a four-wire repeater equipped with pentodes and inverse feed back, and a two-wire repeater also with pentodes and inverse feed back.

In a previous article [1]) published in this Review it has already been pointed out that before the introduction of loading coils the range of transmission in telephony was fairly small. While the adoption of coil-loaded cables increased the range over which calls could be made, it did not entirely remove all restrictions to telephonic intercommunication. An outstanding advance in developing an international telephone service was in fact not made until 1920 when the general principles of amplification technology were worked out.

By inserting loading coils in telephone cables the attenuation can certainly be considerably reduced, but over very great distances the weakening of speech currents is still excessive. For transmission over long distances, it is imperative therefore to supply additional energy to the transmission line by introducing a type of relay action. In the first application of this principle the line was interrupted at a particular point, the incoming line connected to a telephone receiver and the outgoing line to a microphone, the diaphragms of the receiver and the microphone being linked mechanically (Brown relay). The signals reaching the receiver then cause the microphone diaphragm to vibrate, so that an amplified signal is transmitted through the outgoing line. The energy transmitted through the first part of the line is therefore used to operate the relay, while the energy passed through the second part is furnished by the microphone battery.

Nevertheless the general application of amplifiers of repeaters to telephone technique only became possible when the production of the triode amplifying valve provided an inertia-free and efficient relay, which also permitted a high amplification ratio to be obtained.

Two-Wire and Four-Wire Circuits

As stated above the incoming signal is utilised to operate the relay, which in the case of the triode (three-electrode valve) means that the incoming line is connected to the grid of the valve through a transformer. The outgoing line is connected to the anode circuit through another transformer. This circuit will naturally operate in one direction only; to make two-way traffic possible, two methods can be employed, either a two-wire or a four-wire circuit may be adopted (see *fig. 1*).

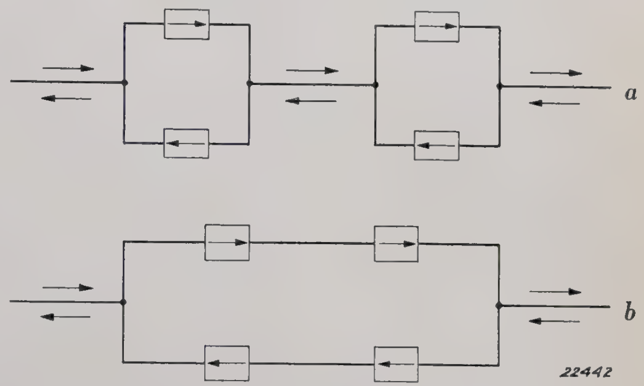

Fig. 1. *a*) General scheme of a two-wire circuit.
b) General scheme of a four-wire circuit.

In the two-wire circuit, two repeaters are connected to the line at the same time, the input of one repeater being in parallel with the output of the second repeater. The repeaters in the various repeater stations are thus linked through a twin-core conductor.

In the four-wire circuit, telephone traffic in the one direction is entirely distinct from that in the opposite direction. Hence two twin-conductors are required and at the terminals the input of one repeater is connected in parallel to the output of the other.

Paralleling the input and output, particularly with two-wire circuits, is not possible without further precautions. For the amplified signals would pass from one repeater to the input of the other repeater, where they could again be amplified, so that in fact the repeaters would commence to oscillate. This is prevented by means of so-called balancing network shown in *fig. 2*, which fundamentally is a bridge circuit. The four arms of the bridge are formed by the line impedance *1*, the artificial line N_2 and the two impedances between the points a - b and b - c; these impedances are determined

[1]) W. Six: The use of loading coils in telephony, Philips Techn. Rev. 1, 353, 1936.

by the input impedance of the repeater. The two transformer windings a - b and b - c are similar, so that the impedances between a - b and b - c also

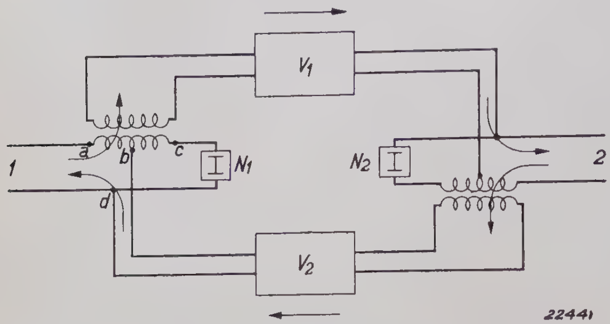

Fig. 2. Balancing network. The impedances of the artificial lines N_1 and N_2 are equal to the impedances of the lines *1* and *2*. In this way the output signal of V_2 produces no voltage at the input of V_1 and vice versa.

are equal. Furthermore, the impedance of the artificial line N_1 is made equal as far as possible to the line impedance for all frequencies. If a potential difference obtains between the points b and d (output voltage of V_2), the potential difference between a and c must be zero, since the bridge is balanced. No voltage is therefore applied to the input of V_1. On the other hand, a signal incoming over the line *1* must be passed to the input of V_1 and after amplification transmitted to the line *2*. But it is difficult to make the impedance of the artificial line exactly equal to the line impedance at all frequencies, since particularly with coil-loaded cables this impedance is non-uniform in a certain frequency range (see *fig. 4*), owing to the irregular intervals between the loading coils and the difference in their self-inductance values.

The propagation of electric waves through cables has already been discussed theoretically in this Review [2]).

From equation (2) of the first article, we can readily deduce that for a wave passing through a cable the ratio $V : I$, which is termed the impedance, is represented by:

$$Z = V : I = \sqrt{\frac{R + j\omega L}{G + j\omega C}}, \quad \ldots \ldots \quad (1)$$

where R is the resistance, L the inductance, G the dielectric conductivity (insulation loss) and C the capacity per unit length of the cable. This expression also applies to coil-loaded cables provided that the frequency is low, i.e. the wavelength is long, compared to the intervals between the loading coils.

But if this is not the case, the fact that the self-inductances are not uniformly distributed over the cable must be taken into consideration. The cable is then more satisfactorily represented by the equivalent circuit shown in *fig. 3*, which is identical to the low-pass filter discussed in the second article. By analogy to the discussion in the series of articles dealing with electrical filters, it may be expected that the attenuation in loaded

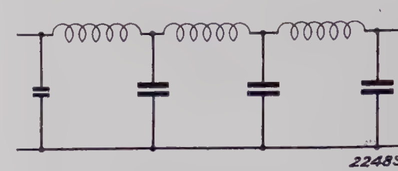

Fig. 3. Equivalent circuit for a coil-loaded cable. If the uniformly-distributed capacity is replaced by condensers, a system equivalent to a low-pass filter is obtained.

cables will remain small up to nearly a specific cut-off frequency of:

$$\nu_1 = \frac{1}{\pi\sqrt{LC}}$$

and will then increase rapidly. Where no losses occur, the impedance, which in the notation used in the second article quoted must be represented by Z_π', should also become infinite at the cut-off frequency and be expressed as a function of the frequency as follows:

$$Z = Z_\pi' = \frac{\sqrt{L/C}}{\sqrt{1 - \frac{\nu^2}{\nu_1^2}}} .$$

These equations for the low-pass filter cannot be applied to cables without some modification, since adequate consideration is not given to the fact that the capacity is distributed uniformly along the line. The qualitative deductions made therefrom are, however, correct, as may be seen from *fig. 4*, where the attenuation and the impedance are plotted

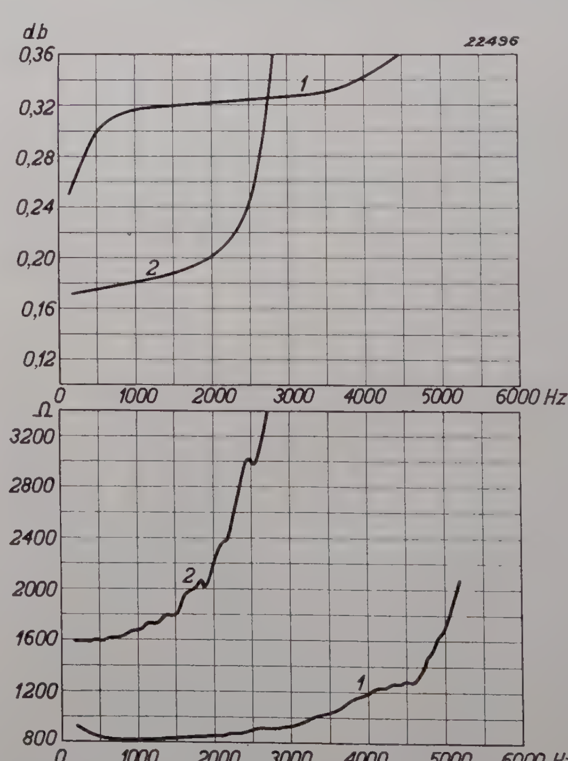

Fig. 4. Attenuation of an impedance in (*1*) a lightly-loaded cable, and (*2*) a heavily-loaded cable.
1 - 44 millihenries at intervals of 1830 m.
2 - 177 millihenries at intervals of 1830 m.

[2]) W. Six: The use of loading coils in telephony, Philips techn. Rev. **1**, 353, 1936, which is the first article; and Balth. van der Pol and Th. J. Weyers, Electrical Filters V, Philips techn. Rev. **1**, 367, 1936, which is the second article.

as functions of the frequency for a low-loaded cable and a high-loaded cable.

Close to the cut-off frequency, the impedance varies considerably even with very small changes in the self-inductance and in the intervals between the loading coils; it gives a wavy curve when plotted as a function of the frequency if the coils are not equally spaced, and for this reason it is difficult to keep the balancing network in balance in this frequency range.

If the bridge circuit is not fully balanced at a specific frequency, then although this circuit will still produce a certain attenuation, part of the signal originating from one repeater will be passed back to the input of the other repeater. As a result the gain is limited, for in the complete circuit which is made up of the two repeaters and the two balancing networks a certain attenuation must be present and must be greater than the total gain in the circuit; if this were not the case the amplifiers would start oscillating.

If the amplification ratio of V_1 and V_2 is, for instance, 20 decibels, then the attenuation in the balancing network must be greater than 20 decibels at all frequencies. But, as already indicated above, this is difficult to achieve close to the cut-off frequency of the cable. A filter is, therefore, connected in front of the amplifiers V_1 and V_2, and serves to produce a marked attenuation of the frequencies lying above the band required for satisfactory acoustic intelligibility. The amplification ratio at frequencies above this band will, in consequence, be much reduced, so that the attenuation of the balancing network can also be made much smaller.

Another important point in the case of two-wire repeaters is with regard to the input impedance, i.e. the impedance between the points *a* and *d*, which must be as nearly equal to the line impedance as possible. The greater the difference between these impedance values, the greater will be the fraction of the incoming signal which is reflected back to the preceding repeater, thus further reducing its stability. Reduction in the stability by reflection can also be described in another way, viz., by saying that the line impedance seen from the points *a* - *d* is determined by the termination of the circuit at the preceding repeater situated at the opposite end of the cable and that therefore the balancing network will only be equivalent when the line is terminated at that point by its impedance.

If, therefore, the impedance of the line is Z and the input impedance of the repeater is W, the coefficient of reflection will be:

$$F = \frac{Z - W}{Z + W} \; ^{3)}.$$

In addition to the input impedance, the impedance of the artificial line must also be equal to Z, as already stated above. *Fig. 5* shows diagram-

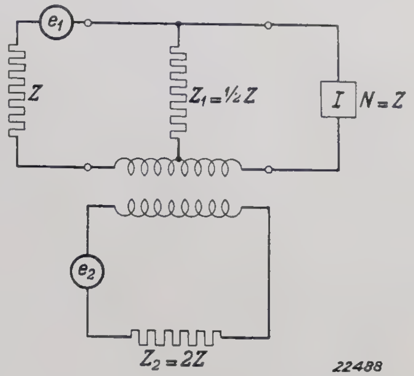

Fig. 5. With this choice of impedances in the balancing network, the bridge is not only balanced but at the same time the line is terminated by its impedance.

matically how these two requirements can be satisfied. In this diagram Z is the impedance of the line, Z_1 the input impedance of the repeater, Z_2 the output impedance of the other repeater and N the artificial line; e_1 is the voltage in the line and e_2 the voltage generated in the amplifying valve. If the impedance of N is equal to Z, the voltage e_2 will not cause current to flow in the branch Z, since the bridge is balanced. If $Z_1 = {}^1/_2\, Z$ and $Z_2 = 2\, Z$, it may be found by calculation that the voltage e_2 produces no current at all through the branch N and that the terminal impedance of the line is equal to Z.

With four-wire circuits, somewhat different conditions obtain. Naturally a certain attenuation must be present also in this circuit which is made up of repeaters, lines and balancing networks. But in this case, the attenuation of the telephone lines also is included in the circuit. If the repeater gain is made so great that the voltage is slightly lower at the end of the line than at the beginning, there can be no question of oscillation, even when a balancing network is not used. The only disadvantage without such a network is that the signal may return to the input of the line and thus give rise to echo phenomena. For this reason balancing networks are used in these cases also, but the artificial lines need not be given the same accuracy in simulation as is required with two-wire circuits.

The amplification ratio at the individual repeater

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3) A cable with impedance can be substituted electrically by a voltage source in series with an impedance Z. If a voltage source e with an internal impedance Z is shorted by an impedance W, the voltage applied to W will be $eW/(Z+W)$. If $Z = W$, the voltage will be $^1/_2\, e$. This may also be regarded as a reflected voltage equal to $^1/_2\, e\,[1 - 2W/(Z + W)] = {}^1/_2\, e\,(Z - W)/(Z + W$.

stations is in this case limited by cross-talk phenomena between the telephone channels as well as by the interference level. In practice a gain of 12 to 15 decibels can be obtained with two-wire repeaters and of 30 to 40 decibels with four-wire repeaters.

Frequency Characteristics of Repeaters

In general, the gain at all frequencies should be made equal to the attenuation produced by the line, i.e. within a frequency band from 300 to 2500 cycles, which as a rule is adequate for telephone channels. With coil-loaded conductors, the attenuation increases considerably at higher frequencies particularly in the neighbourhood of the cut-off frequency. In a heavily-loaded line, i.e. one equipped with coils with high self-inductance values and located close together, the attenuation in the speech band is on the average smaller, but on the other hand the cut-off freqeuncy is lower, while with lightly-loaded lines the attenuation is greater but the cut-off frequency is higher.

It is, therefore, necessary in the case of heavily-loaded lines to insert a wave filter or network between the lines and the repeaters or in the repeaters themselves, whose main purpose is to counteract the increase in attenuation close to the cut-off frequency and at the same time provide a slight balance for the drop in the attenuation at low frequencies. This measure is not necessary with lightly-loaded lines.

Choice of Repeater System

There is still a wide divergence of opinion in different countries with regard to the relative suitability of the various systems for producing a gain in transmission viz., two-wire or four-wire repeaters or light or heavy loading. In Holland the method which has been adopted generally during recent years consists in the use of four-wire repeaters and light loading [5]). Although double the number of conductors are required with the four-wire system, it yet allows a much higher gain to be obtained than the two-wire system, so that a much greater attenuation is permissible in the cables with the same number of repeaters, and in consequence a smaller copper cross-section can be used. Moreover, as already stated above, the four-wire repeater is much simpler to design, since in it the balancing networks with artificial lines are required only at the terminals of the line and need not be made with the same accuracy as in two-wire repeaters; in addition, rejector filters to cut out high frequencies, which are necessary with two-wire repeaters, can be dispensed with in the case of the four-wire type.

The method of loading adopted in Holland in which 65-millihenry coils are located at intervals of approximately 3.68 km, has a cut-off frequency of roughly 3400 cycles, which is sufficiently above the speech frequencies, so that the attenuation in the cable requires compensation only in the range from 300 to 800 cycles. The wide interval between the loading coils has resulted, moreover, in a marked saving in the cost of the coils.

Inverse Feed back

Inverse feed back or negative reaction has been employed in amplifier technique for many years, and consists in feeding back in phase opposition part of the output voltage to the input of the amplifier. In these circuits the gain is naturally reduced, but their advantage lies in the fact that at the same time the distortion is lower and the gain is made more stable, in other words it is made less dependent on the characteristics of the amplifying valve and the feed voltages. For some years inverse feed back has been used in all amplifiers employed in carrier-wave telephony in which an extremely low distortion is essential in view of cross-modulation between the different transmission channels. The Dutch Post Office authorities first employed this circuit in intermediate repeaters for low-frequency telephony, its chief advantage being not so much that the distortion is reduced but that the gain is constant. Reduced gain which is inseparable from inverse feed back is not really a drawback, since by using a valve with a high effective amplification (pentode) roughly the same gain can be obtained as was realised in the past with a triode.

Description of Three Types of Repeater

Three types of repeater have been developed in the Philips Laboratory, all three being designed for indirectly-heated amplifying valves. The use

4) For every section of line joining two repeaters, the average power at the output of one repeater is much greater than that at the input of the other repeater. This difference in level is equal to the gain of the repeater. For lines with opposing directions of speech, the level is high in one line and low in the other line. Cross-talk will therefore be the greater, the greater the difference in level, i.e. the greater the gain.

5) A new telephone system in Holland, by A. H. de Voogt, Post Office Electr. Engin. Journ., **25**, 195, 1932-33.

6) C. J. van Loon: Improvements on radio receivers, Philips techn. Rev. **1**, 264, 1936.

7) A new feed-back repeater, by G. H. Bast and E. H. Stieltjes, Post Office Electr. Engin. Journ. **28**, 225, 1935-36.

of these valves considerably simplifies the circuit compared with triodes having directly-heated filaments as hitherto in use and in which the filaments of four valves were always connected in series. For these latter a separate battery for the negative grid bias was necessary, while at the same time complicated circuits had to be adopted to decouple the repeaters from each other so as to avoid cross-talk. The simplified circuit enables a very compact repeater unit to be obtained. These repeaters are all-mains fed from an alternating-current supply.

I) The first repeater developed was a four-wire repeater with triodes without feed back. A repeater unit, containing four repeaters of this type, is shown in *fig. 6*.

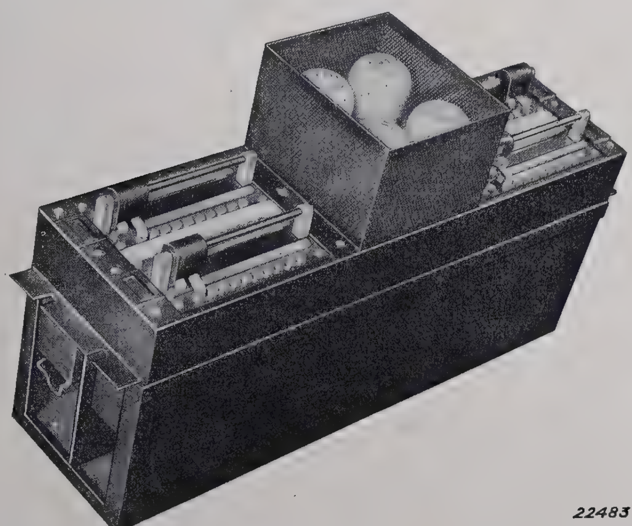

Fig. 6. Repeater unit with 4 four-wire repeaters equipped with triodes.

II) The fundamental circuit of the four-wire repeater with feed back is shown in *fig. 7*. In this repeater a pentode is used as an amplifier, and the feed back voltage is tapped from the rheostat *R*. This voltage is determined by the

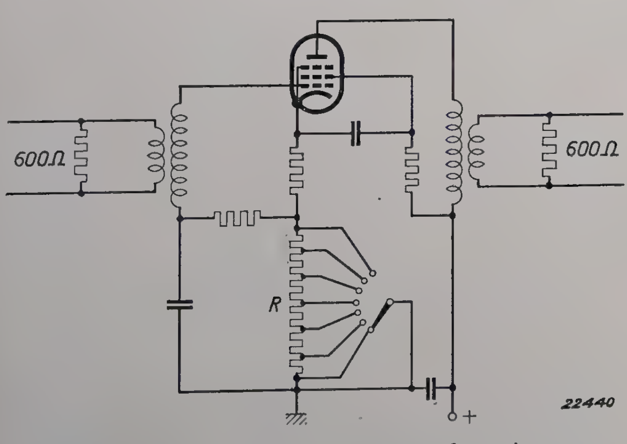

Fig. 7. Fundamental circuit of the Philips four-wire repeater with pentodes and inverse feed back.

resistance, so that the gain is controllable. Between 300 and 4000 cycles the characteristic of the repeater is a straight line within 2 decibels (see *fig. 8*).

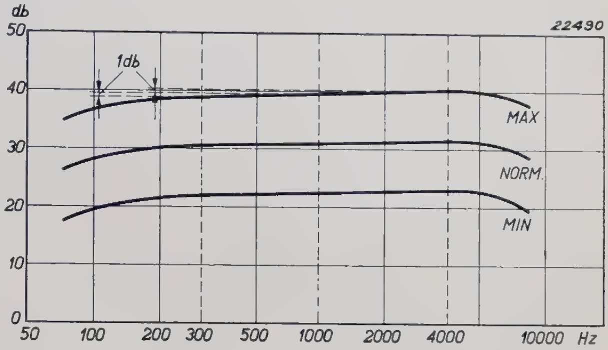

Fig. 8. Gain in the four-wire repeater with inverse feed back plotted against the frequency.

To obtain satisfactory matching to the line, the input and output transformers are terminated with resistances equal to the impedance of the telephone line, viz, 600 ohms. Strictly speaking, the terminal resistance together with the input or output impedance of the amplifier in parallel to it should be made equal to the impedance. The input and output impedances are, in this case, much greater than 600 ohms and hence practically without effect on the combined value. The non-linear distortion for an output of 50 milliwatts is shown in *fig. 9*

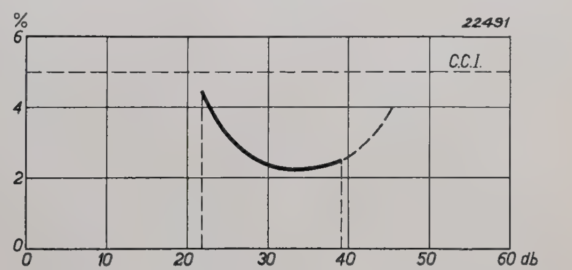

Fig. 9. Non-linear distortion in the inverse feed back repeater plotted against the gain at a power output of 50 milliwatts.

as a function of the amplification. Throughout the whole range this distortion is below the limit stated in the Standards of the Comité Consultatif International des lignes téléphoniques à grande distance (C.C.I). The principal advantage offered by this repeater, as compared with the preceding repeater which is equipped with triodes, is in its marked independence of the feed voltages, as may be seen from *fig. 10*. These curves in fact show that the pentode amplifier with negative reaction still satisfies the C.C.I. standards for lines with more than 12 repeaters when the anode and filament voltages drop to 60 per cent, while in the case of the triode amplifiers the voltage must not drop below 82 per

cent if they are still to conform with these standards.

III) The fundamental layout of a two-wire repeater with inverse feed back is shown in *fig. 11.* In this repeater a combined current and voltage feed back has been used, in other words the negative-reaction voltage is made up of two components: one component is tapped from the resistance *R* and is hence proportional to the output current, and a second component taken from the potentiometer *P* and is hence proportional to the output voltage. The feeding back of the current raises the internal resistance of the repeater, while it is reduced by the voltage back-feeding [8]). By a suitable combination of these two reactions the required internal resistance can be obtained in the repeaters and at the same time the requisite degree of negative reaction realised. In this repeater the internal resistance has been made equal to the line impedance, so that the output transformer need not be terminated by a 600 ohm resistance as is necessary with the four-wire repeater. In the latter, half of the energy output was lost in the terminal resistance, while in the two-wire repeater the whole of the energy output is available for useful work.

In *fig. 12,* the impedance (magnitude $|Z|$ and phase angle φ) of this repeater is plotted against the frequency. The broken line represents the im-

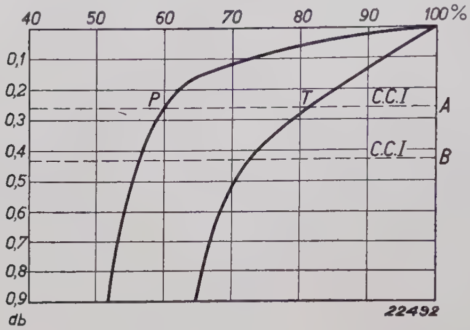

Fig. 10. Gain in decibels below the normal value plotted against the filament and anode voltages in percentages of the normal value. *T* - triode amplifier without negative reaction; P - pentode amplifier with negative reaction. *A* - C.C.I. limits for lines containing more than 12 repeaters. B. - C.C.I. limits for lines with 12 repeaters or less.

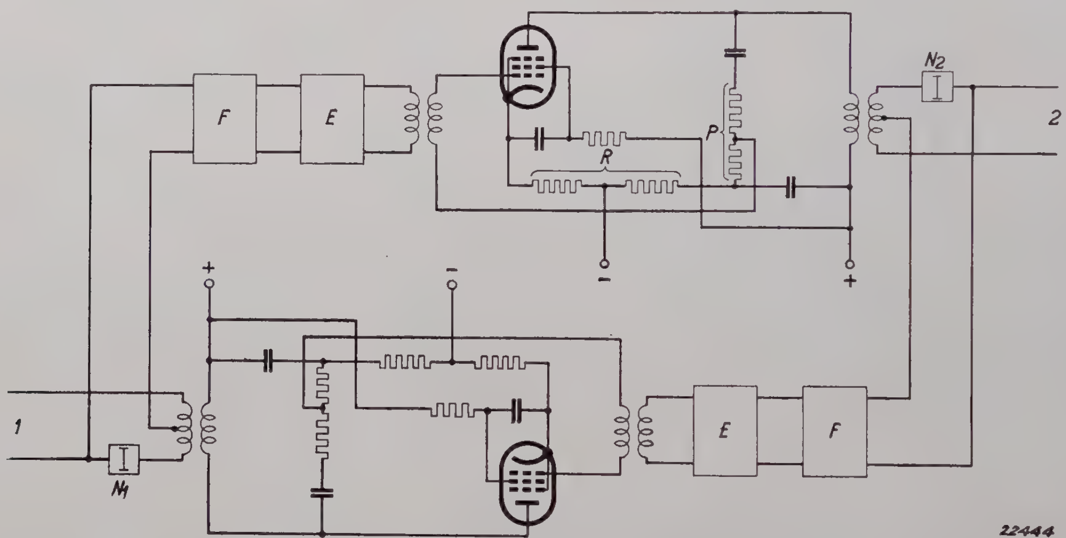

Fig. 11. General circuit diagram of Philips two-wire repeaters with pentodes and inverse feed back. N_1 and N_2 artificial lines. *E* — Balancing networks. *F* — Filters.

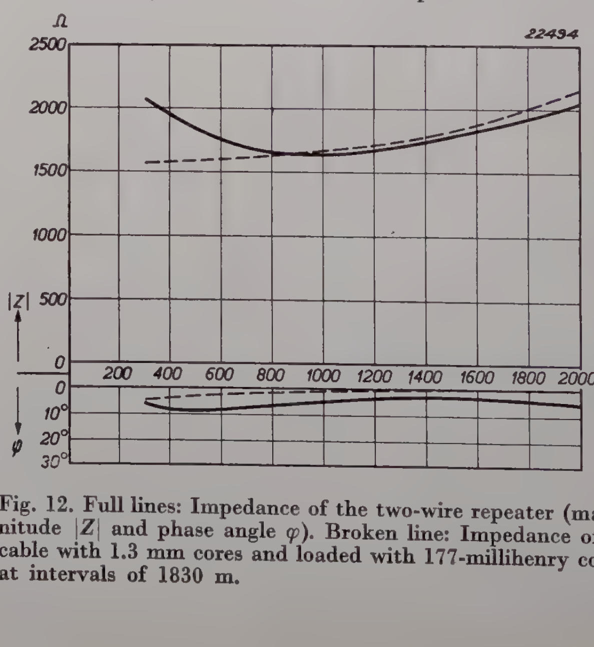

Fig. 12. Full lines: Impedance of the two-wire repeater (magnitude $|Z|$ and phase angle φ). Broken line: Impedance of a cable with 1.3 mm cores and loaded with 177-millihenry coils at intervals of 1830 m.

[8]) Negative reaction or inverse feed back consists in feeding back to the input the voltage variations occurring in one or other of the circuit components at the output side of an amplifying valve, so that these variations become reduced. If the component is so chosen that the voltage applied to it is proportional to the output voltage, the alteration in the output voltage, caused for example by altering the external resistance load connected to it, will be less, i.e. the effective internal resistance is reduced. If the feed-back voltage is made proportional to the output current, this current becomes less dependent on the value of the resistance load, and the internal resistance is increased.

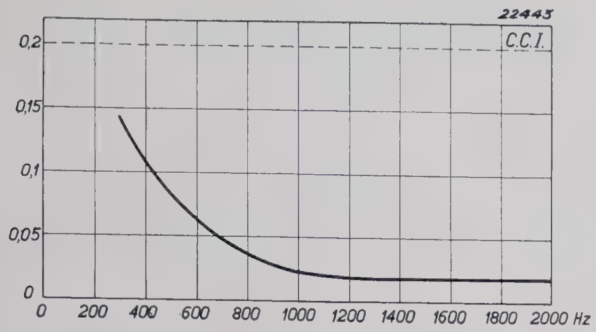

Fig. 13. Coefficient of reflection of the two-wire repeater plotted against the frequency, when using a cable for which this repeater has been designed.

pedance of the cable for which the repeater was designed (diameter of cores 1.3 mm; loading coils of 177 millihenry at intervals of 1830 m).

The reflection coefficient $F = Z - W/Z + W$ is plotted against the frequency in *fig. 13*. Throughout the whole frequency range this coefficient is considerably below the limiting value specified by the C.C.I.

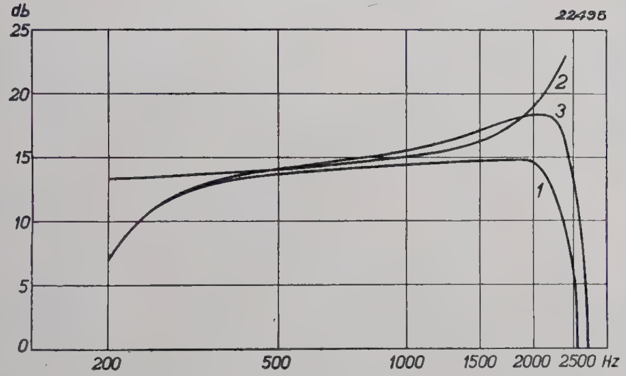

Fig. 14. Gain in two-wire repeater containing pentodes, plotted against the frequency. *1* - Repeater with rejector filter for high high frequencies. 2 - Attenuation of the associated cable. 3 - Repeater corrected by means of a balancing network.

Fig. 14 gives the frequency characteristic of this repeater. Curve *1* is the characteristic of the repeater with rejector filter for cutting out high frequencies. Curve *2* is the attenuation of the cable, while curve *3* is the characteristic of the repeater when using a balancing network which compensates for the increase in attenuation in the neighbourhood of the cut-off frequency of the cable.

Finally, a photograph of a repeater of this type is reproduced in *fig. 15.*

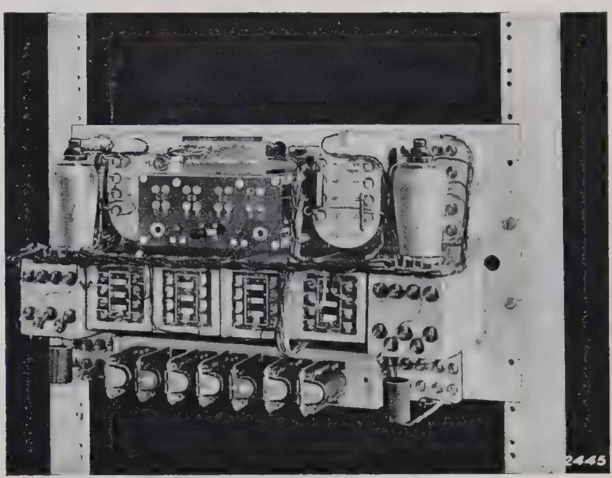

Fig. 15. Photograph of a Philips two-wire repeater with pentodes and inverse feed back.

A RECORDING FIELD-STRENGTH METER OF HIGH SENSITIVITY

by M. ZIEGLER.

Summary. An apparatus for measuring the signal strength of radio transmitters radiating on wave lengths between 10 and 2000 m is described. The sensitivity of the apparatus is calibrated during measurement by means of an oscillator incorporated in it. Field strengths from approximately 1 microvolt per m upwards can be measured with this apparatus.

Introduction

The measurement of the field or signal strength of radio transmitters is now a matter of routine in wireless engineering and may be carried out for a variety of practical purposes. A few applications in which the field-strength meter has proved indispensable may be quoted from the innumerable uses to which this instrument has already been put: The energy radiated from a transmitting aerial with a known input is a criterion for determining the efficiency of the aerial, while the distribution of field strength may be explored over the area surrounding a transmitter to ascertain the intensity of interference at a particular point, due either to an interfering transmitter or to an oscillating receiver. The directional characteristics of an array of transmitting aerials can be investigated, as well as the absolute minimum and maximum signal strengths due to fading ascertained. The purpose of a measurement of the field strength may have purely a scientific interest, but in the majority of cases measurements have primarily a technical value, such as those enumerated above. Practical data concerning the power radiated from a specific type of aerial, as well as a close knowledge of the effects of local geological and topographical conditions on the propagation of waves, are essential for planning and selecting the best site for a transmitting station which shall operate efficiently and be economical to construct and run.

The various problems entailed in a measurement of the field strength are very diverse, and the practical and theoretical requirements imposed on the measuring apparatus used are widely divergent. Firstly, very weak fields due to remotely-situated transmitters must be capable of measurement, and secondly, measurements must be feasible in the immediate vicinity of the transmitting aerial; in other cases again the apparatus must be used on sites to which it cannot be conveniently transported by vehicles. For still other purposes, it may be desirable to install the measuring arrangements permanently in order to carry out measurements at a particular spot of the variations in signal strength with time over periods of several weeks. Furthermore, on certain occasions measurements have to be carried out on long waves and on other occasions on ultra-short waves.

Every type of measurement of the field strength required in normal practice can be carried out directly with the apparatus described below. Before passing to a description of the apparatus, the principle on which the field-strength meter operates must be briefly outlined.

Principle of the Field-Strength Meter

A field-strength meter is composed of an aerial in which a certain electromotive force is induced by the field of the transmitter, and a voltmeter for measuring this e.m.f.

The effective height of the aerial is of particular interest in this connection, and is defined as follows: If the electric field strength for a particular wave is $e = E \sin \omega t$, then an e.m.f. of

$$V = h\,e \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

will be induced in an aerial of effective height h. If V is measured and the effective height is known, the signal strength can be calculated. The effective height of a vertical aerial with a high top capacity (*fig. 1a*) and whose length l is small as compared

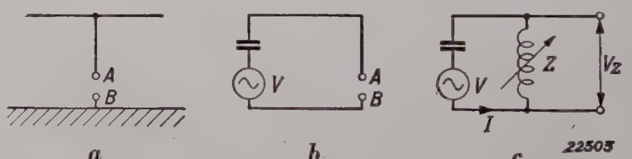

Fig. 1. *a*) Vertical aerial with high top capacity (capacity aerial). A vertical electric field induces a voltage V which is equal to the product of the field strength and the length of the aerial.
b) Equivalent circuit of the capacity aerial. The induced e.m.f. may be regarded as being connected in series with the capacity of the aerial.
c) By connecting a self-inductance Z in series with the aerial, a circuit tuned to the impressed signal can be obtained. At resonance the voltage V_Z can be many times V.

with the wave length is l for a vertically-directed electric vector; the e.m.f. induced by the wave will then be $V = l\,e$.

Adopting the above definition it is also possible to speak of effective height in the case of a frame aerial. The effective height of a frame aerial (flat

coil, *fig. 2a*) of n turns, each enclosing a surface O whose dimensions are small as compared with the wave length, is

$$h = \frac{2\,\pi\,n\,O}{\lambda}\sin\alpha, \quad \cdot \cdot \cdot \cdot \cdot \quad (2)$$

where α is the angle between the plane of the frame and the magnetic vector. As is well known a frame aerial has directional characteristics. A frame aerial suitable for use with e.g. a portable field-strength meter has 16 turns of $0.4 \cdot 0.4 = 0.16$ sq. m surface, when for $\lambda = 300$ m and $\alpha = 90$ deg, we get $h = 0.054$ m. This height is considerably less than the effective height of a vertical aerial of several meters.

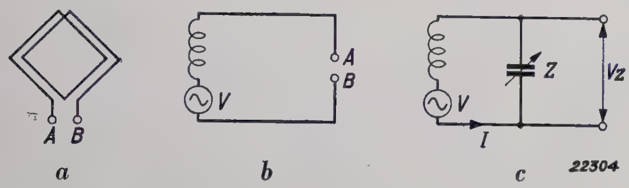

Fig. 2. *a*) Frame aerial (inductive aerial). A magnetic field of unit intensity, which is perpendicular to the plane of the frame aerial, induces a voltage $V = \frac{2\,\pi\,n\,O}{\lambda}$ where n is the number of turns, O the area of the frame, and λ the wave length.
b) Equivalent circuit of the inductive aerial. The induced e.m.f. may be regarded as connected in series with the self-inductance of the frame aerial.
c) By connecting a capacity Z in series with the frame aerial, a circuit tuned to the received signal can be obtained. At resonance the voltage V_Z can be many times V.

The equivalent circuits of the capacity aerial and the frame aerial with an assumed induced e.m.f. are represented in figs. 1*b* and 2*b* respectively.

The problem therefore resolves itself into a measurement of the voltages V indicated in figs. 1b and 2b. But only the terminals A and B are accessible. Theoretically, it is sufficient to measure the potential difference between A and B with a voltmeter, whose impedance is high compared to the reactance of the vertical aerial or the frame aerial. Although it is in fact not imperative, the resonance principle will be used in this case, and a circuit tuned to the signal under measurement formed by connecting an inductance or a condenser in series, the selective and amplifying characteristics of this circuit offering important advantages (figs. 1*c* and 2*c*). Either the current I which flows through the circuit can be measured, or the potential difference V at the series-connected impedance. If at the same time the resistance of the circuit is ascertained, V can be directly calculated, for:

$$V = Ir = V_Z\,\frac{r}{Z}\,.$$

What type of aerial is provided with the field-strength meter?

The requirements of an aerial for use in conjunction with the field-strength meter are that it shall have a specific ascertainable effective height and at the same time be transportable. A frame aerial automatically satisfies these requirements, while a capacity aerial would assume an unfavourable form, for instance those mounted on a short vertical pole fitted with a horizontal plate at the top would have a capacity with respect to earth which is high as compared with that between the vertical part and earth. It is thus reasonable to consider a condenser aerial, but the effective height of an aerial of this type cannot readily be made appreciably greater than that of a standard frame aerial.

A further difficulty with the capacity aerial is that it cannot be so easily tuned as a frame aerial, for a variable selfinductance is required for this purpose (fig. 1*c*) which cannot be realised to suit all cases. In the more normal circuit shown in *fig. 3* the gain in the voltage due to the field is

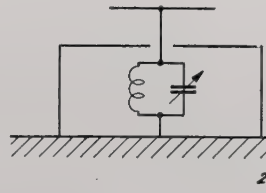

Fig. 3. A short vertical pole on which a horizontal plate is mounted, is an example of a mobile capacity aerial of well-defined characteristics. The effective height is equal to the distance between the plate and the top of the metal box enclosing the measuring apparatus.

much smaller than the resonance amplification obtainable with the frame aerial. Since, moreover, a frame aerial has directional characteristics as compared with a capacity aerial, which in general will prove an advantage, as a standard it is to be preferred to a condenser aerial.

However, where an aerial with a great effective height is essential for the purpose of measuring very weak signals, and if the directive action is either dispensable or even undesirable, one or other of the capacity aerials may be employed. Any aerial system, whose effective height is unknown and which is coupled in any manner to an amplifier with an indicating instrument (e.g. an aerial installed in a motor car), can always be calibrated as a whole:

1) By comparison with a standard aerial, and
2) By measuring the theoretically-known field of an accurately-calibrated testing transmitter.

Important requirements which the field-strength meter must meet

A field-strength meter has to meet the following requirements:

a) High sensitivity in order to measure weak signals,

b) High selectivity to permit weak signals to be measured even when interference is produced by more powerful stations on an adjoining frequency;

c) Constant characteristics, in order to avoid the need for frequent recalibration in the laboratory.

To meet requirements a) and b) a sensitive amplifier is used containing a number of amplifying valves in cascade and various tuned circuits, by means of which the induced voltage is amplified for measuring purposes. On the other hand the characteristics of the apparatus cannot be made constant over long periods by any simple and direct method, for the sensitivity varies with the age of the valves and is to a great extent determined by the accuracy of tuning, etc. For this reason, the apparatus has been so designed that the sensitivity can be adjusted to the required value as necessary.

The sensitivity of the apparatus can be checked very easily, viz, by observing the effect of a known e.m.f. which has the same frequency as the field under measurement and which is connected in series with the frame aerial as indicated in *fig. 4*. Not only can the sensitivity of the voltmeter, which in this case includes the amplifier and the instrument, be checked by this method, but also the gain which occurs in the resonance circuit formed of the frame aerial and the condenser. *The calibration voltage produces exactly the same effect as an induced e.m.f. of the same magnitude* [1]), and will therefore also give the same deflection on the voltmeter. If the induced e.m.f. V under measurement is not equal to the calibration voltage, the deflection of the voltmeter will be reduced or increased in the corresponding ratio. By this comparison method the reliability of measurements made is mainly dependent on the accuracy of the calibration voltage.

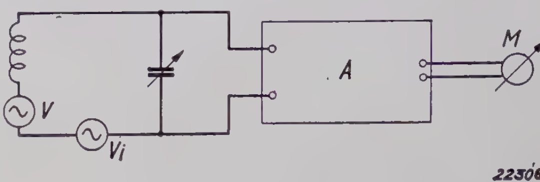

Fig. 4. Principle of the field-strength meter. The voltage V induced in the frame aerial is compared with the calibration voltage V_i. To measure low voltages, the voltage which is produced by V or V_i at the condenser is amplified by an amplifier A.

Electrical Layout of the Apparatus

In fig. 4 the general principle of the Philips field-strength meter is illustrated and in *fig. 5* a detailed layout of the various component circuits. Referring to this diagram we shall first investigate how the signal under measurement produces a deflection on the measuring instrument.

The voltage of frequency ν which is induced at the terminals of the aerial circuit, as already described above, is passed through a selective amplifying stage H and then to the mixer valve M. The intermediate-frequency signal of frequency ν_m which is produced by interaction with the auxiliary voltage of frequency $\nu + \nu_m$ originating from the oscillator O_h, is filtered out and is passed through a variable attenuator V of which further details are given below. Subsequently, the signal is amplified

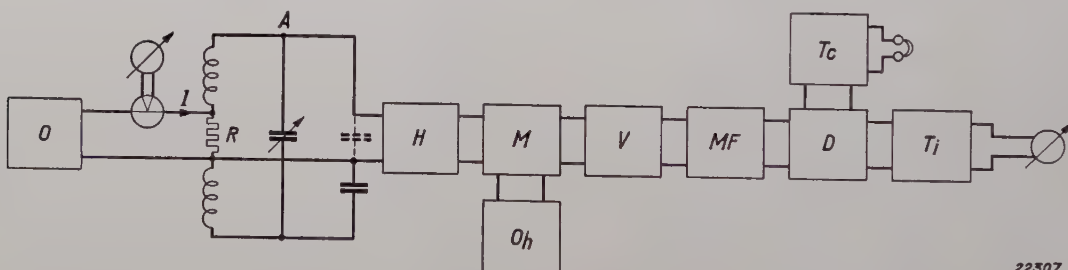

Fig. 5. Diagrammatic layout of the signal-strength meter. O Calibration oscillator to furnish the calibration voltage $V_i = IR$. A Aerial circuit; H High-frequency amplifier; M Mixing valve; O_h Heterodyne oscillator; V Attenuator; MF Intermediate frequency amplifier; D Detector; T_i Indicator triode, which furnishes the current for the instrument; T_c monitor triode for the current passed to the headphones.

[1]) This is not quite correct, since the frame aerial is never entirely devoid of capacity and the induced e.m.f. is not concentrated at a single point. But deviations from these ideal conditions can be neglected in practical measurements.

by two stages MF tuned to the frequency ν_m and then rectified in a diode detector D. The direct voltage now produced is passed to the grid of a triode T_i of which the anode current flows through a milliammeter. This triode, which we shall term the indicator valve, is so adjusted that its anode current is 5 milliamps when no signal is impressed on it. A rectified signal makes the grid of the triode negative with respect to the cathode, so that the anode current drops and produces a deflection of the pointer from the no-signal reading. The total amplification of the signal under measurement must be high enough to give a deflection easy to read, yet not so high that the anode current of the indicator valve becomes zero. The amplification stage is also controllable in addition to the attenuation stage, by means of the mixer valve and the two intermediate-frequency valves, so as to permit accurate adjustment.

The ratio between two signals of the same frequency can be directly read off on the instrument without altering the adjustment. This ratio cannot be determined if it is greater than 10, but with the attenuation stage, the amplification of one signal can be reduced or increased in a definite ratio as compared with the amplification of the other. The maximum attenuation is 1000, so that voltages in a ratio up to 1 : 3000 can still be compared with each other.

Naturally, at all practical adjustments, the amplification ratio must be independent of the amplitude. Both the maximum amplification in each stage and the method by which the amplification is varied are so calculated that the maximum deflection is obtained on the instrument before any other part of the apparatus becomes overloaded.

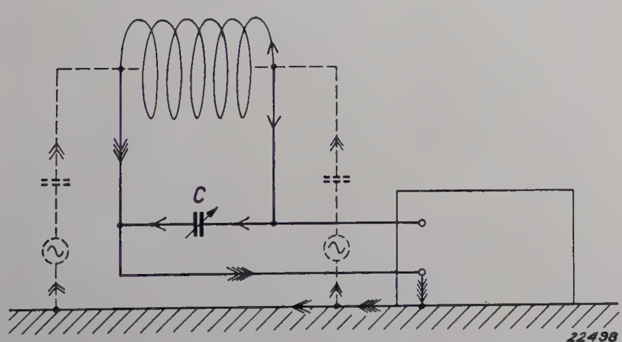

Fig. 6. Asymmetrical frame-aerial circuit. If one end of the aerial winding is earthed, the electric field will cause currents to flow, which, independently of the direction of the frame aerial, will produce a voltage at the terminals of the amplifier. The arrows indicate the currents which are produced by the vertical electric field between the frame aerial and earth. It is seen that part of these currents flow through the condenser C, and hence produce an alternating voltage at the input terminals of the amplifier.

We shall now give a description of the various components of the field-strength meter.

Aerial circuit: Frame aerial. If a frame aerial is connected up as shown diagrammatically in fig. 4, the capacity of the aerial and that of the apparatus with respect to earth (*fig. 6*) will, as a whole, exhibit the same characteristics as a capacity aerial, so that a different type of signal will be generated that would be deduced from equation (2). This follows from the fact that the signal does not fall to zero, even when the plane of the frame is parallel to the magnetic vector. This undesirable characteristic, the socalled "aerial effect", may be eliminated by mounting the frame symmetrically, when the currents generated by the capacity effect balance each other (*fig. 7*).

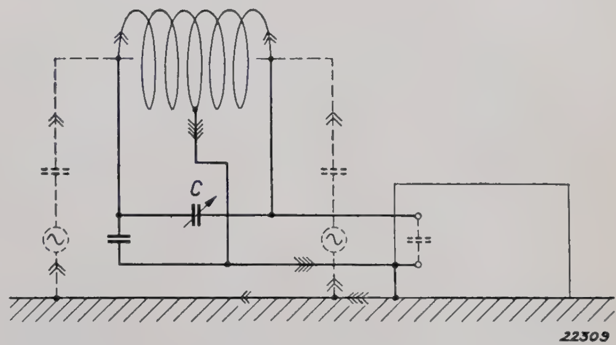

Fig. 7. Symmetrical frame-aerial circuit. Since the centre of the aerial winding is earthed, the currents induced by the electric field counteract each other. No current flows through the condenser C.

The frame aerial is tuned with a variable condenser C. The measuring equipment is provided with a series of interchangeable frame aerials.

Reception with capacity aerial. For reception with a capacity aerial, the rotatable frame holder can be replaced by coils which automatically make the necessary contacts, so that a circuit of the type shown in *fig. 8* is obtained. Each coil is fitted with

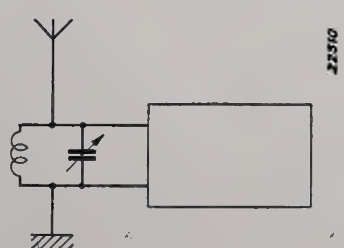

Fig 8. Connection of the aerial circuit when using the signal-strength meter with a capacity aerial.

a terminal to which the aerial is connected; the apparatus is earthed.

The calibration signal. The potential difference at the resistance R (fig. 5) is used for calibration and

is produced by an alternating current I of the required frequency furnished by a special oscillator (O). As already indicated in fig. 4, the calibration voltage is in series with the frame aerial. The resistance is only of the order of 1/8 ohm, so that for $I = 8$ milliamps, $V_i = 1$ millivolt. The sensitivity of the thermo-couple which is used for current measurement and the resistance R are independent of the frequency, hence also the calibration voltage. The oscillator through its five adjustable ranges furnishes all frequencies between 150 and 30 000 kilocycles and is so rated that the maximum current flowing through the thermo-couple will never overload the couple. This maintains the characteristics of the thermo-couple constant and also raises the reliability of the apparatus as an absolute measuring instrument.

High-frequency Amplifier. The function of this amplifying stage, which consists of an amplifying valve with tuned anode circuit, is twofold, viz:

1) To raise the high-frequency selectivity in order to suppress those signals with a frequency $\nu + 2\nu_m$ which may give rise to a disturbing intermediate-frequency signal, and
2) To make the ratio of the signal strength to the interfering noises as satisfactory as possible. By using a high-frequency amplifying valve with very low mush the noise level can be reduced to the unavoidable thermo-electric fluctuations in the first circuit. Even when using a small frame aerial a signal due to a field of only a few microvolts per m will still be sufficiently audible above the mush.

Mixer Valve and Auxiliary Oscillator. An ordinary mixer hexode is used as a mixing valve, and a special triode which has five switched wave-length ranges similar to the calibration oscillator is connected up as an oscillator. The tuning condenser is coupled mechanically to the condenser in the high-frequency stage, so that for tuning the apparatus only one other knob in addition to that for the aerial condenser has to be adjusted. The gain in the mixer valve is regulated by means of the variable negative grid bias.

Attenuator. The attenuator V is coupled to the mixer valve by a pair of circuits tuned to the intermediate frequency signal. By means of a switch the attenuation factor can be adjusted as required to 1, 3, 10, 30, 100, 300 or 1000. The attenuator consisting of wound wire resistances which have been very carefully calibrated, is balanced so that the input and output resistances have exactly the same values at all adjustments. The attenuated signal passes to the first intermediate-frequency amplifier through two coupled tuning circuits.

Intermediate Frequency Amplifier. This amplifier has the following components and characteristics: Two amplifying valves, four tuned circuits, and controllable gain by variation of the negative grid bias. In addition, the grid bias of both valves can be automatically controlled by the impressed signal, such that strong signals are amplified less than weak signals, which may be desirable when recording signal strengths subject to marked fluctuation.

Rectification. The amplified intermediate frequency voltage is passed to the two diodes of a duo-diode-triode. One of the diodes serves exclusively for generating a direct voltage for the automatic gain control just referred to, the whole or part of this voltage being applied to the grid of the intermediate frequency amplifiers. Rectification by the other diode furnishes a low-frequency alternating voltage with a D.C. voltage component, the latter being passed to the indicator valve described above. The alternating voltage is passed to the triode grid of the duo-diode-triode. Reception can be followed audibly by means of headphones connected to the anode circuit of this triode.

Indicating Instrument. The milliammeter with a range of 5 milliamps, which gives readings of the anode current, is connected up in such a manner that the pointer is to the right of the scale when no current is flowing. When a current of 5 milliamps flows the pointer will give a reading to the left of the scale opposite the scale zero. In the absence of a signal, the anode current can be adjusted to the correct value by regulating the potential difference between the cathode and the grid. The scale has been calibrated by experiment in such a way that with a constant sensitivity, i.e. when not using the automatic gain control, the deflection on the instrument is proportional to the input signal strength.

This division of the scale is determined principally by the I_a-V_g characteristic of the triode, so that it is important for the operating characteristics of this valve to remain constant. The anode current is, therefore, maintained automatically constant by means of a neon lamp and is thus unaffected by any fluctuations in the voltage supply. A second instrument, e.g. a recorder, can be connected in series with the milliammeter incorporated in the apparatus.

The field-strength meter with frame aerial and battery box is shown in *fig. 9*.

Variable Selectivity. Although the meter has ten circuits, the selectivity has not been made excessively sharp, since the eight intermediate-frequency circuits are connected in pairs as band filters. The coupling of two pairs of these circuits has been made variable, so that the selectivity can be increased by making the coupling looser.

Current Supply. The apparatus is designed for connection to either a direct-voltage supply (batteries) or a 50-cycle A.C. mains supply.

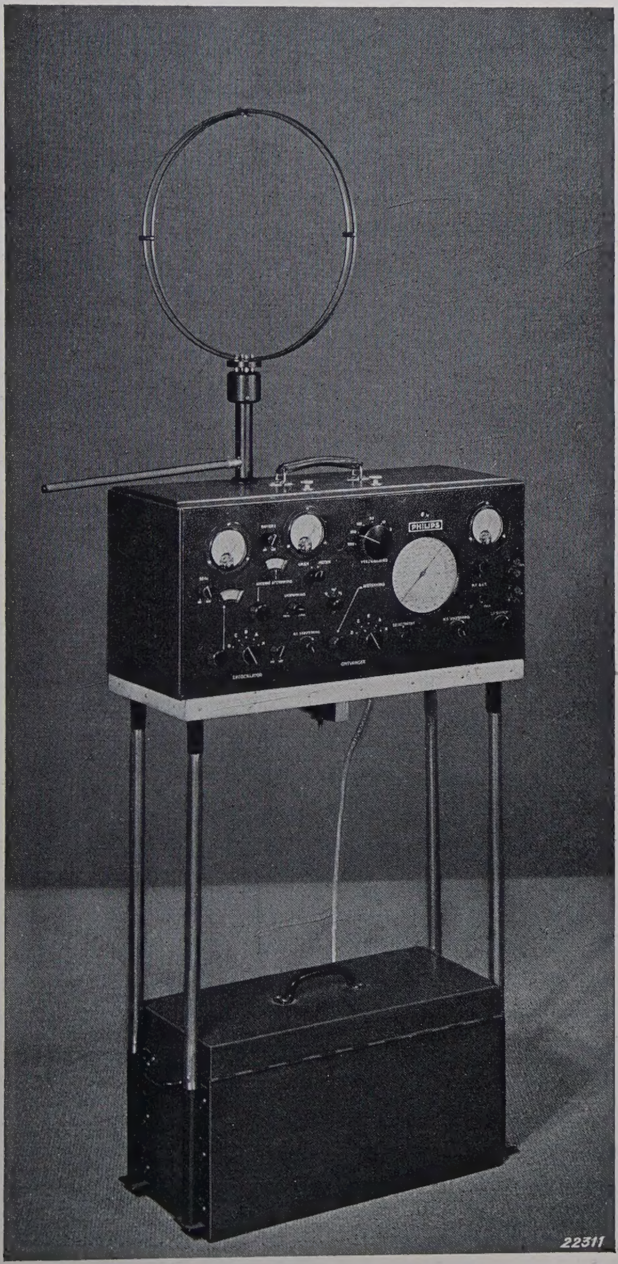

Fig. 9. Philips' field-strength meter with battery box. The apparatus is equipped with a frame aerial for wave lengths from 28 to 75 m. On the left of the front panel is the meter for the calibration voltage; on the right is the meter which indicates the output current. The large circular tuning scale is provided for frequency calibration in each of the five frequency ranges. Above and to the left of the tuning scale is the knob for the attenuator.

Exemples of various measurements

The application of the field-strength meter is illustrated by the following practical examples of its use;

a) *Field-Strength of a Local Transmitter*. One of the most important uses of the field-strength meter is for measuring the field distribution in the area about a transmitting station. The apparatus described here is particularly suitable for this purpose, since it is capable of measuring signal strengths over a very wide range of values.

Assume that the signal strength of the Hilversum station, transmitting on a wave length of 300 m with an aerial output of 15 kilowatts, has to be measured in the neighbourhood of Muiden.

As already stated above, the effective height of the frame aerial provided for this wave length, when suitably orientated, is 0.054 m at maximum signal strength. By accurate tuning and satisfactory orientation of the frame aerial, and using an attenuation of e.g. 1/30, without automatic gain control, the pointer can be adjusted to read 5 by regulating the amplification. The frame is now turned into the position corresponding to the minimum signal strength (there will remain a residue of e.g. 1 per cent), and the calibration oscillator is switched in. After tuning and adjusting the current at the low resistance to 8 milliamps (calibration voltage = 1 millivolt), a reading of 6.5 is obtained with an attenuation of 1/10, all other adjustments remaining unaltered. Hence the effective value of the signal originating from the field radiated by Hilversum is

$$V_{\text{eff}} = 1 \cdot \frac{30 \cdot 5}{10 \cdot 6.5} = 2.3 \text{ mV}.$$

The effective value of the field strength is thus

$$\frac{V_{\text{eff}}}{h} = \frac{2.3}{0.054} = 42.2 \text{ mV/m}.$$

Where the field strength is to be explored over a large area, the field-strength meter can be installed in a motor van. It is then an advantage to use, in place of the frame aerial, a vertical aerial 0.5 to 1 m in height, which has no directional characteristics at all and which does not have to be taken down when moving from place to place. The absolute calibration of the mobile field-strength meter is carried out with a field strength previously explored by a frame aerial by the method described above.

b) *Signals radiated from radio receivers*. It is essential to suppress as far as possible the radiation of signals from auxiliary oscillators in superheterodyne receivers in view of the mutual interference

which they cause in reception. In a particular case, it was stipulated that the field radiated by a receiving set equipped with a particular aerial should not exceed 5 microvolts per m at a distance of 25 m.

To investigate whether a particular set satisfies this condition on a wave length of 30 m, a frame aerial is used which is adapted to this wave length and has two annular windings of 40 cm diameter. The effective height in the optimum position is 0.0525 m. The field radiated by the receiver under investigation is so small that the calibration signal is much more powerful than the induced signal, so that the frame aerial before calibration does not have to be turned into the position for which the received signal has a minimum strength.

The amplification is so adjusted that a reading of 10 is obtained with a calibration voltage of 500 microvolts and 1/1000 attenuation.

After switching off the calibration signal, a deflection of only 3.1 is found for the induced signal without attenuation. The signal strength is therefore

$$V_{\text{eff}} = 500 \cdot \frac{3.1 \cdot 1}{10 \cdot 1000} = 0.155 \ \mu\text{V}.$$

and the effective value of the field strength is:

$$E_{\text{eff}} = \frac{0.155}{0.0525} = 2.95 \ \mu\text{V/m},$$

c) *Recording the signal strength of a distant transmitter.*

The transmission of electromagnetic waves over very great distances has been the subject of systematic investigations during recent years. *Fig. 10* reproduces an oscillogram of the signal strength of the L.R.1. transmitter at Buenos Aires, which was recorded at Eindhoven during the night hours.

For this measurement a vertical aerial 12 m in height was used, which owing to the lack of top capacity had an effective height of only 6 m. The measuring equipment was supplemented by a recording milliammeter. As the automatic gain control was in circuit, the deflection of the pointer increased more slowly than the intensity of the received signal; this is shown in fig. 10. In cases where the field strength is expected to show very large variations, the amplitude can be controlled still more effectively than was done in the present case.

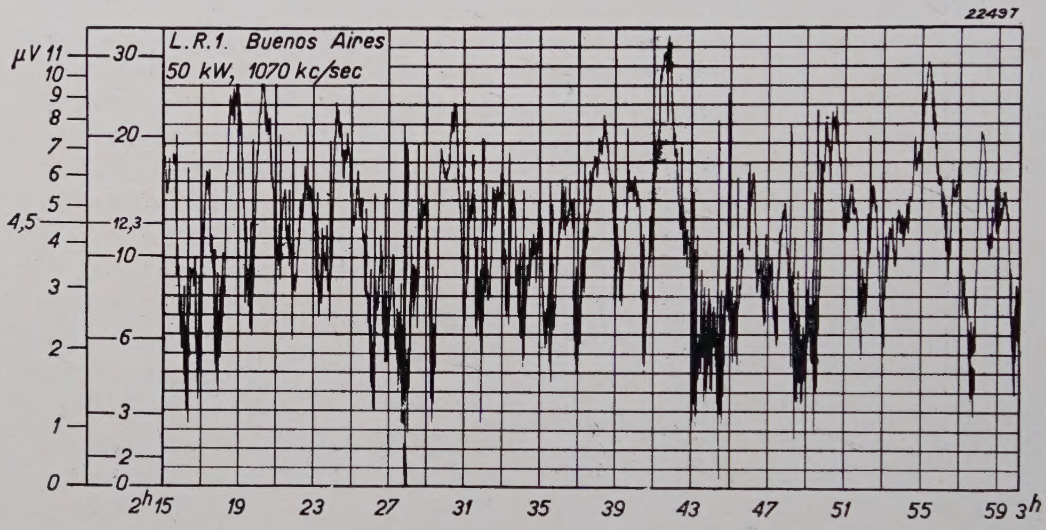

Fig. 10. Oscillogram of the field strengths of the L.R.1. transmitting at Buenos Aires, recorded on December 15, 1935, at Eindhoven. The scale 0 to 30 was obtained by plotting the deflection for calibration signals of 1 and 0.5 millivolt using different settings of the attenuator. A station with a field strength of 450 micro-volts per m and about the same frequency thus gives a deflection of 12.3 = 4.5 microvolts per m when using an amplification 100 times smaller than employed in the according; this deflection gives the absolute calibration of the scale.

The division of the relative scale from 0 to 30 was carried out with a constant calibration signal and the attenuator, by keeping the adjustment of the automatic gain control unaltered and varying the intermediate frequency signal in the ratios of 1, 3, 10, 30, etc. and then determining the deflection on the oscillogram. The scale value in microvolts per m was determined by picking up a station transmitting on approximately the same frequency and having a constant field strength, the latter being measured by the method described under a) above.

If no interference is experienced in reception, field strengths of less than 0.1 microvolt per m can

be measured by this method. In the recording strip reproduced, the interference level lay between 1 and 2 microvolts per m.

Use of the Apparatus as a Voltmeter

In conclusion, reference should be made to an important use to which this meter can be put, viz., as a selective voltmeter with adjustable sensitivity and high aperiodic input impedance for measuring sinusoidal alternating voltages. After removing the frame aerial or aerial coil, the alternating voltage to be measured is applied to the grid and cathode of the first amplifying valve, using the special terminal provided for this purpose. With a switch the unknown voltage V_x can be compared to the voltage V_i of the calibration oscillator (*fig. 11*). The sensitivity of the instrument as voltmeter is approximately 1 millivolt.

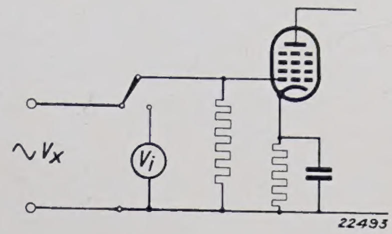

Fig. 11. Circuit using the meter as voltmeter. The unknown voltage V_x is compared to the voltage V_i of the calibration oscillator.

REVIEW OF RECENT SCIENTIFIC PUBLICATIONS OF THE N.V. PHILIPS' GLOEILAMPENFABRIEKEN

No. 1172: M. J. Druyvesteyn and N. Warmoltz: Ein neuer Dunkelraum in der Nähe einer Glühkathode in einer Bogenentladung (Physica 4, 51 - 68, Jan. 1937).

A new dark space round the oxide-coated cathode of an arc discharge, which has already been briefly described in Abstract **No. 1117,** is discussed in detail in this paper. This dark layer is about 20 times the thickness of the space charge sheath close to the cathode and occurs at current densities above 0.1 amp per sq cm and at pressures between 10^{-2} mm and 3 mm. At lower pressures the new dark sheath disappears and a bright sheath with the same dimensions appears in its place. Starting from the cathode, a sharp increase of the concentration and a decrease of the mean energy of the fast electrons occur at the boundary between this layer and the bright part of the discharge. This may be accounted for by a scattering of the fast electrons which leave the boundary of the space charge sheath close to the cathode. This scattering increases with increasing current. A dark or bright sheath is observed according as the increase in concentration of the fast electrons outside this layer or the reduction in mean energy of these electrons has the greater effect on the emission of light.

No. 1173: K. F. Niessen: Über die Wirkung eines vertikalen Dipolsenders auf ebener Erde in einem Entfernungsbereich von der Ordnung einer Wellenlänge (Ann. Phys., **28,** 209 - 224, Jan. 1937).

Formulae which are sufficiently accurate for application up to distances of the order of a wave length are deduced for the intensity of the horizontal component of the magnetic field and the vertical component of the electric field of a vertical dipole transmitter. These formulae are then simplified for types of soil and wave lengths for which the Sommerfeld numerical distance is small compared with unity.

No. 1174: H. C. Hamaker: A general theory of lyophobic colloids. II. Rec. Trav. chim. Pays Bas, **56,** 3 - 25, Jan. 1937).

Continuing the analysis reviewed in Abstract **No. 1154,** a formula is deduced which gives the total energy of interaction between two colloidal particles, expressed approximately as a function of the distance apart of the particles; it also contains as independent parameters the charge E of a colloidal particle and the electrolyte concentration **c.** In an E - c diagram the different states of a colloid can be represented as a function of E and c. A variety of phenomena, such as peptisation, reversible and irreversible flocculation and thixotropy, can be analysed in a clear and simple way by means of the curves in the E - c diagram.

No. 1175: J. A. M. van Liempt and P. Leydens: Die Farbenwiedergabe beim Photographieren mit Neonlicht (Rec. Trav. chim. Pays Bas, **56,** 26 - 28, Jan. 1937).

Colour reproduction in photography with neon light is investigated for different negative materials;

it is found that neon is a satisfactory source of light for orthochromatic materials. However, owing to the shorter exposure times, preference is usually shown for panchromatic material, with which good results may be obtained by using a mixture of neon and mercury lights. Nevertheless, the quality of this mixture is slightly less satisfactory than the mixture of sodium and mercury lights (cf. Philips Techn. Rev. **2**, 24, 1937).

No. 1176: J. A. M. van Liempt and J. A. de Vriend: Studien über das Verbrennungslicht einiger Metalle (Rec. Trav. chim. Pays Bas, **65**, 126 - 128, Jan. 1937).

The quantity of light emitted during combustion, as well as the comparative efficiencies of the combustion process is determined for tungsten, molybdenum, tantalum, cerium and carbon.

No. 1177: M. J. O. Strutt: Les performances de certains types des lampes changeuses de fréquence dans les récepteurs toutes ondes (Onde él. **16,** 29 - 44, Jan. 1937).

The various requirements which mixing valves for short waves have to satisfy are summarised under eleven headings. The most important characteristics of octodes operating on short waves are discussed and the means indicated for overcoming the difficulties encountered. The characteristics of octodes in general are discussed and the latest improvements in construction are briefly stated. In this connection, data are given of the behaviour of the latest types of valves for short waves, which show that they are improvements on the types in use hitherto.

No. 1178: J. H. Gisolf: The counting of electrons by means of a discharge tube (Physica, **4,** 69 - 70, Febr. 1937).

Different types of discharge tubes can be used as electron counters in a special circuit, which is described in detail in this paper.

No. 1179: F. M. Penning: Ein neues Manometer für niedrige Gasdrücke, insbesondere zwischen 10^{-3} und 10^{-5} mm. (Physica **4,** 71 - 75, Febr. 1937).

A glow-discharge tube in a magnetic field may be adapted for the measurement of gas pressures, especially between 10^{-5} and 10^{-3} mm. The current passing between two electrodes is a useful measure of the pressure. The length of the negative glow on the rod-shaped cathode of another glow-discharge tube is a very suitable indicator for this current. This new manometer, which is marketed by E. Leybolds' Nachfolger A.G., Cologne-Bayenthal, under the name of the Philips' Vacuummeter, is very simple in its use. It gives an uninterrupted, instantaneous and clearly visible indication of very low pressures, those of condensable vapours included. (Cf. also the description of this instrument on p. 201 of this issue).

No. 1180: J. Sack: The charge of electricity on metal drops which leave the welding rod during electric-arc welding (Physica, **4,** 104 - 106, Febr. 1937).

Liquid drops of metal, which during welding pass from the welding rod to the work, are caught before reaching the latter in order to measure their charge with a Wulf string-electrometer, and in addition their weight is determined. The drops from the coated iron welding rod PH-50 always carry a negative charge, both in A.C. and in D.C. welding, irrespective of whether the polarity of the welding rod is positive or negative. This negative charge of 2 to $3 \cdot 10^{-9}$ coulomb is relatively high for such small drops of metal of only 3 to 4 mm diameter, and can produce a voltage of 10 kV below earth. A satisfactory explanation of this phenomenon has not yet been given.

No. 1181: A. Bouwers, F. A. Heyn and A. Kuntke: A neutron generator (Physica **4,** 153 - 159, Febr. 1937).

In this paper, a tube is described by means of which high-speed ions and neutrons can be produced. A high-voltage generator used in connection with this tube is also discussed, as well as a new method for supplying the potential to the ion-source.

No. 1182: F. A. Heyn: The radio-activity of cobalt, nickel, copper and zinc induced by neutrons (Physica **4,** 160 - 165, Febr. 1937).

The radio-activity of cobalt, nickel, copper and zinc induced by neutrons of different energies is investigated. If copper and zinc are bombarded with fast neutrons, half-life values are obtained which suggest the existence of a new nuclear reaction in which two neutrons are emitted by the bombarded nuclei. Experiments to confirm this assumption are described.